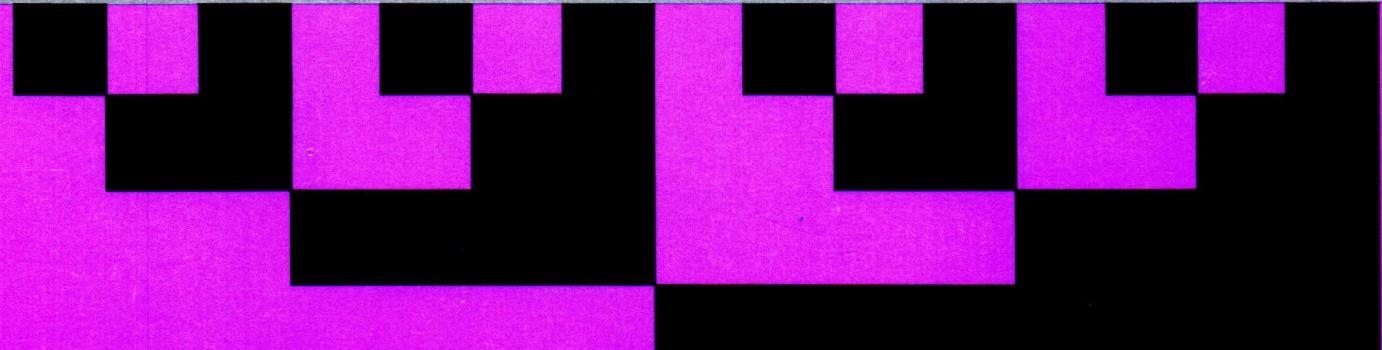




PHILIPS

digital instrument course

Part 3 DIGITAL VOLTMETERS AND MULTIMETERS



22.429

mV

V

kΩ

MΩ

mA

°C

Published by

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**The front cover shows the pattern of the BCD 1, 2, 4, 8
code and the digital display of a voltmeter.**

DIGITAL INSTRUMENT COURSE

Part 3. Digital voltmeters and multimeters

by M. H. van Erk and H. G. Onstee

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Part 1 Basic binary theory and logic circuits

The first part of the digital instrument course deals with the fundamentals of binary theory and logic circuitry.

The chapter headings are:

Number systems

Boolean algebra

Logic elements

Combinational logic

Bistable elements (flip-flops)

Counters, scalers and shift registers

The circuitry of logic elements

Interfaces and measuring systems

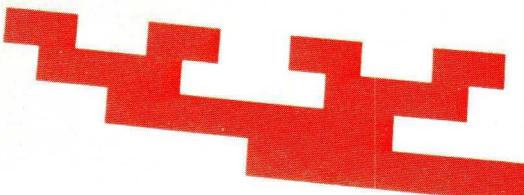


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Part 1

BASIC BINARY THEORY AND LOGIC CIRCUITS



PHILIPS

digital instrument course

Part 2

DIGITAL COUNTERS AND TIMERS



123456789. Hz

Part 2 Digital counters and timers

The second part of the digital instrument course deals with digital frequency counters and timers

The chapter headings are:

Basic counter circuitry

Modes of operation

Plug-in units and special functions

Accuracy

Foreword

This book is part three of our digital instrument course, based on a training course initially set up for instrument specialists in our European sales organizations.

The first part covers basic binary theory and logic circuits, while part two deals with digital frequency counters and timers.

The present book – digital voltmeters and multimeters deals with the principles of operation of DVM's, and highlights two application-oriented techniques involving DVM's.

A digital voltmeter (DVM) measures analog voltages and presents the result as a decimal figure. The multifunction version of this instrument, the digital multimeter (DMM), also measures currents and resistances.

Compared with analog voltmeters and multimeters, the digital instruments have opened two new dimensions in measuring techniques. First of all, the accuracy. In analog displays (in most cases with a moving-coil meter), the accuracy was limited mainly by the length of the scale and the ability to read the position of the pointer on the scale correctly. Use of a decimal display does away with this resolution problem, as we can extend the "scale" to any desired number of decimal digits. In this way full use can be made of all advances in precision technique, and DVM's with up to 6 digits are now used widely.

A second aspect of these DVM's is that the conversion of analog information into a digital form makes it possible to transport the data over a long distance, to process them in a computer, store them and compare them accurately with preset limits, thus opening the way to automation.

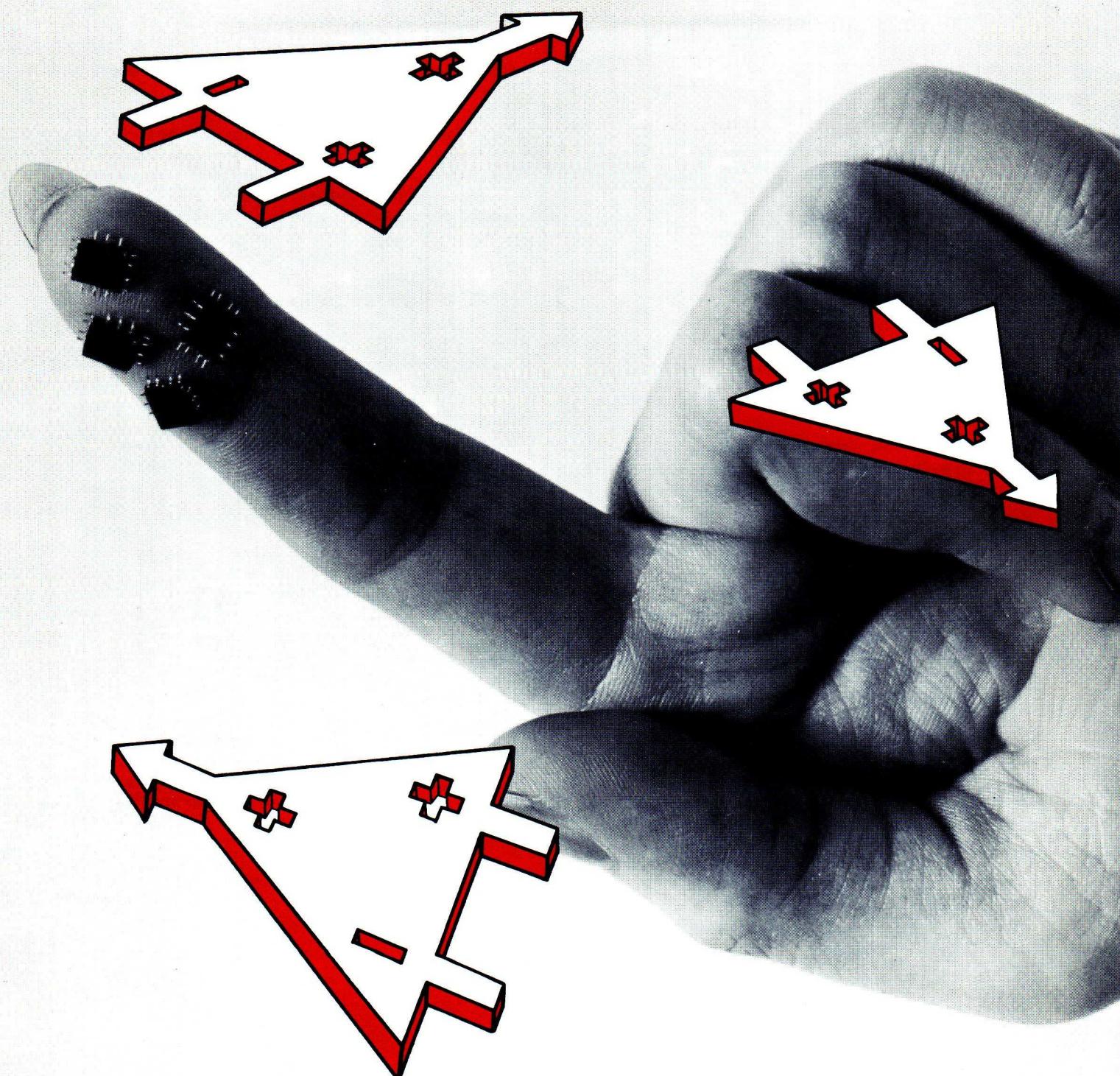
Although the first DVM's were rather bulky and moreover very expensive, they soon became popular. Interaction between growing demand for these instruments and technological developments gradually formed the DVM and the DMM into the popular low-priced precision instruments they are today.

Digital voltmeters still contain some analog circuits in addition to the purely digital sections. In particular, operational amplifiers ("opamps") perform several vital functions in the analog part of digital voltmeters used for processing the input signal and it is essential to understand these functions before discussing the rest of the circuitry. Chapter 1 therefore gives some basic information on operational amplifiers. The essential part of each DVM or DMM is still the analog-to-digital converter (ADC), which will be discussed in chapter 2. Automatic functions such as autoranging, autozeroing and automatic polarity indication are a logical implementation of the ease with which the digital information can be used for various purposes, and will be dealt with in chapter 3. The analog circuitry that conditions or converts the incoming signal to the DC voltage needed by the ADC will receive attention in chapter 4.

Last but not least, we shall see in chapter 5 how the extremely high accuracies obtainable with a DVM can be maintained thanks to a knowledge of the error sources involved.

With the improvements in the performance of the DVM in terms of higher sensitivity, resolution and accuracy, the effect on measurements of parasitic signals from the surroundings came to play a more and more important role and special measures had to be taken to ensure reliable results. One of these measures is proper guarding and shielding. The techniques involved are dealt with in chapter 6.

The modern components of today make it a simple matter to perform true RMS measurements instead of conventional AC measurements, in which the waveform of the signal is assumed to be sinusoidal. These techniques and their possibilities are described in the last chapter 7. We trust that this book will help you to gain the understanding of DVM's you need for your work. It goes without saying that we would welcome your criticism and comments.



Chapter 1

The operational amplifier

Basic principles

An operational amplifier is an amplifier with a very high open-loop voltage gain. This gain G can amount to over 100 dB ($= 10^5$), and even reach levels of 10^7 – 10^9 .

The most widely used version of the operational amplifier in voltmeters is the one with differential input and single-ended output, the symbol for which is shown in fig. 1.1.

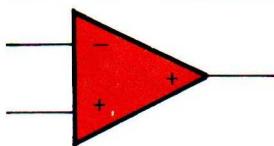


Fig. 1.1. The symbol for an operational amplifier with differential input and single-ended output

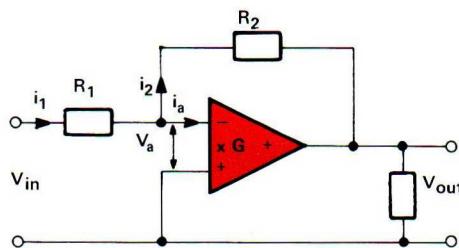


Fig. 1.2. An operational amplifier used with a feedback loop. G is "open loop" gain of the amplifier

This type of amplifier can be used for several purposes. Thanks to its very high gain, it is very suitable as a sensitive zero detector and can be used in combination with a suitable feedback loop as a preamplifier, integrator, AC-DC converter, impedance transformer or constant-current source, all functions often used in digital voltmeters and multimeters.

It is obvious that without the availability of inexpensive IC operational amplifiers the DVM and DMM never would have achieved such a widespread acceptance.

To understand the function of the operational amplifier in its many application we will first have a look at its behaviour when used with a feedback loop as in fig. 1.2.

The input marked with a – is called the "inverting input", as any signal applied to this will appear with opposite polarity at the output.

As it will be proved later, the input voltage V_a is very small and as the input impedance of the op.amp. itself can be quite high we may assume that the input current of the op.amp., i_a , is nearly zero.

As a result we state that $i_1 = i_2$

$$\text{As } i_1 = \frac{V_{in} - V_a}{R_1} \text{ and } i_2 = \frac{V_a - V_{out}}{R_2}$$

we may write

$$\frac{V_{in} - V_a}{R_1} = \frac{V_a - V_{out}}{R_2}$$

$V_{out} = -GV_a$ (G is the open-loop gain, and we are operating with inverted polarity).

$$\text{so } V_a = -\frac{V_{out}}{G}$$

Substituting this in above formula we find

$$\frac{V_{in}}{R_1} + \frac{V_{out}}{GR_1} = -\frac{V_{out}}{GR_2} - \frac{V_{out}}{R_2}$$

and since G is very high it follows that

$$\frac{V_{in}}{R_1} \approx -\frac{V_{out}}{R_2}$$

and the closed-loop gain

$$\frac{V_{out}}{V_{in}} \approx -\frac{R_2}{R_1}$$

We see that the closed-loop gain depends only on the ratio of R_1 and R_2 , thanks to the high value of the open-loop gain G .

$$\text{As } V_a = -\frac{V_{out}}{G}$$

V_a is very small (e.g. $V_{out} = 10$ V and $G = 10^8$ gives $V_a = -0.1 \mu\text{V}$); we may thus say that the inverting input of this amplifier is "virtually earthed". This also means that the input impedance of the whole circuit is $\frac{V_i}{i_1} \approx R_1$, and hence completely independent of the other parts of the circuitry.

Applications of the operational amplifier in voltmeters

We now will have a look at some basic voltmeter and multimeter circuits in which operational amplifiers are used. Some of them are used in the analog circuitry and we will meet them again when discussing the analog parts. The others are found in the analog-to-digital converters dealt with in the next chapter.

Comparator or zero detector

This application of an op.amp. is easy to understand. The signals to be compared (V_A and V_B) are applied to the inputs of the operational amplifier, fig. 1.3.

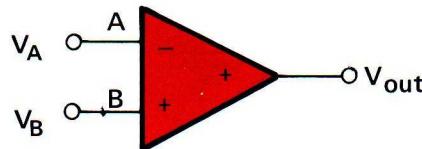


Fig. 1.3. Operational amplifier used as a comparator or zero detector

Because of the high gain of the amplifier a very small difference in input signal (of the order of $0.1 \mu\text{V}$) will result in a full excursion of the amplifier's output voltage; so if V_A is only some tenths of a microvolts lower than V_B the output will be fully positive and in the opposite case (V_A slightly higher than V_B) the output will be fully negative. When V_B is connected to earth the circuit compares V_A with zero, giving a sensitive zero detector which changes its output polarity when V_A passes zero with an error of only a small fraction of a microvolt.

Preamplifier with adjustable gain

In the formula

$$\frac{V_{\text{in}}}{R_1} \approx -\frac{V_{\text{out}}}{R_2}$$

the full output voltage is used for feedback. When only a fraction $\frac{R_4}{R_3 + R_4}$ of the output signal is used for feedback (see fig. 1.4), the formula becomes

$$\frac{V_{\text{in}}}{R_1} \approx -\frac{\frac{R_4}{R_3 + R_4} V_{\text{out}}}{R_2}$$

The closed-loop gain is thus $\frac{V_{\text{out}}}{V_{\text{in}}} \approx -\frac{R_3 + R_4}{R_4} \cdot \frac{R_2}{R_1}$

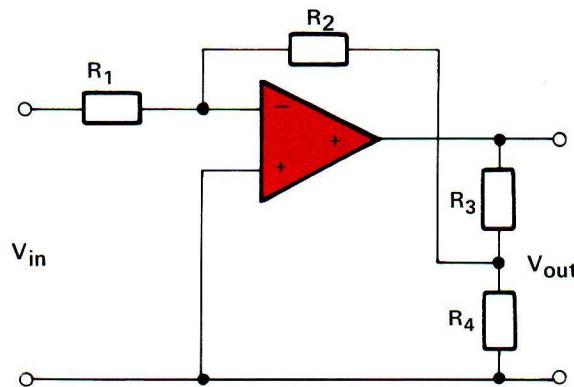


Fig. 1.4. Operational amplifier used as an adjustable gain pre-amplifier

Attenuation of the output voltage before feedback thus gives a higher gain. This has the special advantage that low-ohmic attenuators (of the order of $1 \text{ k}\Omega$) can be used, with less risk of noise and hum.

Constant-current source

As we have seen above,

$$i_2 = i_1 = \frac{V_{\text{in}}}{R_1}$$

So if we keep V_{in} and R_1 constant, the current i_2 will also have a constant value, that can be set by the relation between V_{in} and R_1 . This circuit has the advantage that very low constant currents can be generated; a disadvantage is that the load has to be floating, fig. 1.5.

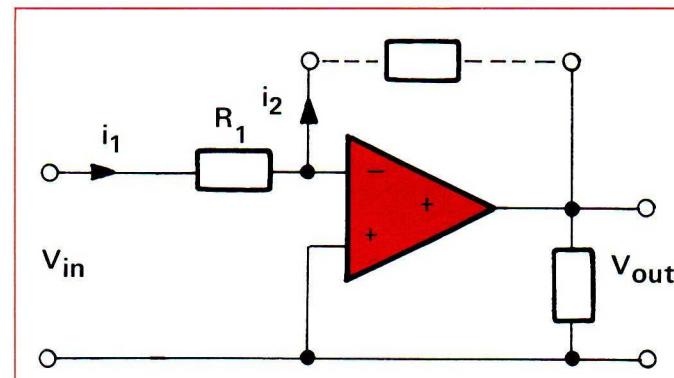


Fig. 1.5. Operational amplifier used as a constant-current source

Integrator

With a capacitor in the feedback loop, the op.amp. can be used as an integrator, fig. 1.6.

As long as the capacitor C in the feedback loop is not fully charged, it acts as a DC path and does not change the characteristics of an op.amp. with feedback loop.

As a result

$$i_2 = i_1 = \frac{V_{in}}{R_1}$$

Because the inverting input is virtually earthed (V_a is very small compared with V_{in}), the output voltage V_{out} is equal to the negative voltage across the capacitor V_c .

So

$$-V_{out} = V_c = \frac{Q_c}{C} = \frac{1}{C} \int i_2 dt = \frac{1}{C} \int \frac{V_{in}}{R_1} dt.$$

or

$$-V_{out} = \frac{1}{R_1 C} \int V_{in} dt.$$

Questions:

Check the answer(s) you believe to be correct.

Q.1.1. An operational amplifier is an amplifier with

- A a very low input impedance
- B a very high open-loop gain
- C a resistor in the feed-back loop

| | |
|---|--|
| A | |
| B | |
| C | |

Q.1.2. In the circuit of figure 1.7. the closed-loop gain $G = \frac{V_{out}}{V_{in}}$ is independent of the gain of the operational amplifier because:

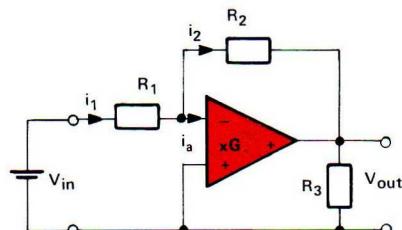


Fig. 1.7.

- A the input current i_a is zero
- B the open-loop gain G is very high
- C R_2 is connected in the feed-back loop

| | |
|---|--|
| A | |
| B | |
| C | |

Q.1.3. In figure 1.7. the closed-loop gain is

- A R_3/R_2
- B R_2/R_1
- C R_1/R_3

| | |
|---|--|
| A | |
| B | |
| C | |

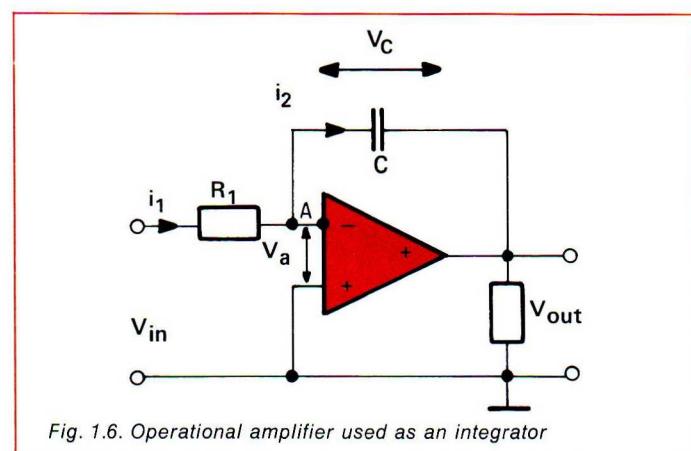


Fig. 1.6. Operational amplifier used as an integrator

or in words:

The output voltage of the operational amplifier is directly proportional to the integrated input voltage V_{in} .

Q.1.4. The circuit of figure 1.7. may be considered as a constant-current source for the current through R_2 because:

- A V_{in} and R_1 are kept constant
- B R_2 is large compared to R_1
- C the input current i_a is zero

| | |
|---|--|
| A | |
| B | |
| C | |

Q.1.5. The circuit of figure 1.8. is

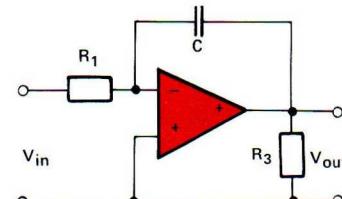


Fig. 1.8.

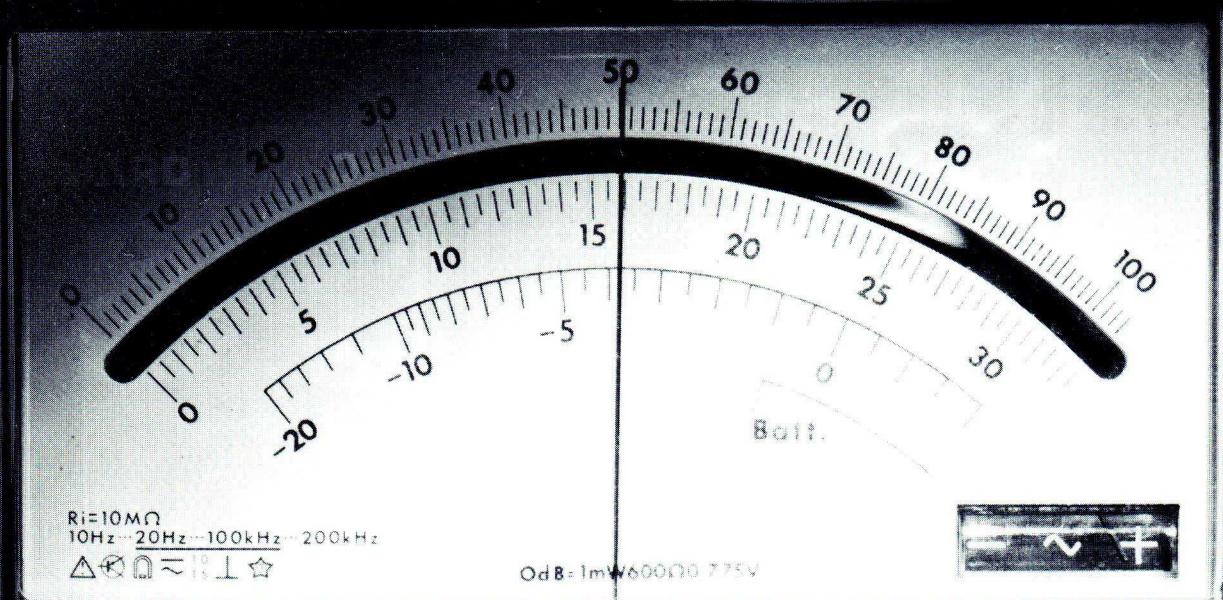
- A an integrator
- B a differentiator
- C an adder

| | |
|---|--|
| A | |
| B | |
| C | |

The answers to the problems will be found at the end of the book.

digital rms multimeter

03.94



PM 2504 electronic VAΩ meter

POWER

ON

100

Chapter 2

The analog-to-digital converter (ADC)

Function of the ADC

The function of the analog-to-digital converter (ADC) in a voltmeter is to convert the analog input signal (in nearly all cases a DC voltage) into a discrete number of "steps" that can easily be counted and presented in decimal form on a numerical display.

The steps into which the analog input signal is converted may be of different natures, e.g. voltage, time or charge, depending on the type of ADC involved.

In all cases, however, such a step should correspond to an accurately known voltage.

A good way of determining the value of the step involved is the standard method of comparing it with an accurately known reference voltage.

If the step is a DC voltage, this is quite a simple matter. When the step is of another nature (time, charge, etc), the comparison can be made by converting the reference voltage into the same steps with the aid of the same circuit or an identical one.

This means in fact that in all accurate digital DC voltmeters, the voltage to be measured is compared with a reference voltage and the relation between input voltage and reference voltage is what counts.

Basic principles of the ADC

There are several different principles on which an ADC can be designed. The most widely used systems are:

Direct compensation

The input signal is compared with an internally generated voltage, which is increased in steps starting from zero. The number of steps needed to reach full compensation is counted.

Voltage-to-time conversion

Both input voltage and reference voltage are converted into a time and the ratio of the two times is counted.

Voltage-to-frequency conversion

The input voltage is converted into a number of pulses and the number of pulses in a well defined constant time is counted.

In this chapter we will discuss some of the basic circuits used for ADC's in greater detail.

First of all, however, it is useful to consider what are the most important points of any analog-to-digital converter used in digital voltmeters and multimeters.

Requirements on ADC's in voltmeters and multimeters

There is no single set of requirements which guarantees optimum performance of ADC's in all different types of voltmeters and multimeters. The aspect to which highest priority should be given depends on the application. For this reason, the order in which the requirements are mentioned below is not necessarily the order of importance for all applications.

Accuracy

One of the advantages of a digital instrument is the very high resolution it can give, which can permit a very high accuracy of appreciable sources of error can be avoided. The over-all accuracy is mainly determined by the ADC, so the number of sources of error in this stage must be kept to a minimum to ensure high accuracy and stability.

Series-mode rejection

All ADC's described here measure DC voltages. However, the voltage presented to the input of the ADC is rarely a pure DC voltage. In nearly all cases, it also contains AC components such as hum and noise.

Now we want the ADC to measure the DC component only. The AC signal (the series-mode signal) should thus be rejected. A filter could be used to reject this signal but has the disadvantage of slowing the instrument down. Another solution is to integrate the signal over a certain period of time. If the integration period is long enough the average AC signal will be nearly zero and a good series-mode rejection will be obtained (see also the section on "sources of error" in chapter 5).

Speed

For normal bench operation, 2 to 3 measurements per second are quite sufficient, but 10–50 or more measurements per second may be required for applications in automatic measuring systems. The speed at which the ADC operates can be limiting in these cases.

Kick-back signals and input impedance

The more accurate the measurement the more important it is that the voltmeter itself does not influence the signal to be measured. We must thus make sure that the input impedance is high and that feedback of signals from the voltmeter to the signal source is really minimum.

(See also the section on "Sources of error" in chapter 5)

At the end of this chapter we will compare the degree to which the above requirements are fulfilled by the different types of ADC's.

The staircase ramp; a simple compensation system

The basic principle of the circuit of fig. 2.1. is that the input signal V_i is compared with an internal staircase voltage V_c generated by a series circuit consisting of a pulse generator (clock), a counter counting the pulses and an digital-to-analog converter (DAC) converting the counter output into a DC signal.

As soon as V_c is equal to V_i the input comparator closes a gate between clock and counter, the counter stops and the counter output is shown on the display.

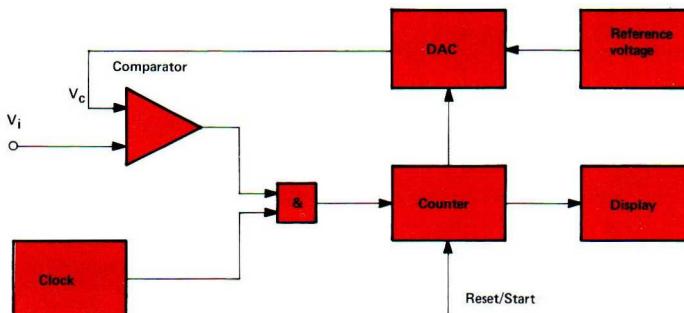


Fig. 2.1. Block diagram of a direct-compensation ADC

The operating sequences in this circuit is as follows:

The clock generates pulses continuously. At the start of a measurement (time t_1) the counter is reset to zero so that the output of the DAC is also zero (for a description of the operating principles of the DAC see chapter 8 in Part 1 of this course).

If V_i is not equal to zero, the input comparator supplies an output voltage that opens the gate so that clock pulses are passed on to the counter. The counter starts counting and the DAC starts to produce an output voltage increasing by a small step at each count of the counter. The result is a staircase voltage supplied to the second input of the comparator (see figure 2.2).

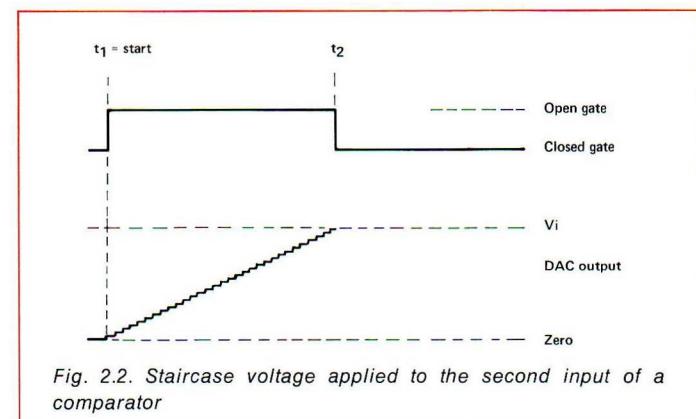


Fig. 2.2. Staircase voltage applied to the second input of a comparator

This process continues until the staircase voltage is equal to or slightly greater than the input voltage V_i . At that moment (t_2) the output voltage of the input comparator changes in polarity, so the gate closes and the counter is stopped. The display unit shows the result of the count. As each count corresponds to a constant DC step in the DAC output voltage, the number of counts is directly proportional to V_c and hence to V_i . By appropriate choice of the reference voltage on which the step height of the staircase voltage depends, each count can represent e.g. 1 mV and a direct reading of the input voltage in e.g. volts can be realized by placing a decimal point in front of the 10^3 decade.

The system can be reset by applying a new reset/start pulse to the counter.

The advantages of this system are:

- Thanks to the compensation the input impedance of the ADC is high when compensation is reached.
- The accuracy depends only on the stability and accuracy of the reference voltage and the DAC. The clock frequency has no influence on the results as the time in which compensation is reached does not affect the accuracy.

The disadvantages are:

- The system measures the instantaneous value of the input signal at the moment compensation is reached. This means that the reading is rather unstable if the input signal is not a pure DC voltage (i.e. the system has rather poor series mode rejection).
- As long as full compensation have not yet been reached the input impedance is low, which can influence the accuracy.
- The kick-back signal is rather high because of changing input conditions during the measurement and the risk of cross-talk from clock pulses to the input.

Successive-approximation method

The successive-approximation ADC works along much the same line as the staircase ramp system just described. The main difference is that the compensation signal is built up more quickly by starting with higher steps as long as the difference between input signal and compensation signal is high.

Figure 2.3. shows a simplified block diagram of this type of ADC.

At the start of a measurement the counters are reset so that the output of the DAC will be zero. The chopper at the input is alternately connected to V_i and V_c , so when an input signal is applied the chopper output will be a square wave with an amplitude depending on the difference between V_i and V_c (see figure 2.4.).

The pulse shaper converts the positive slope of this square wave into pulses with an amplitude related to that of the square wave and this signal is amplified by three amplifiers in series. The resulting signals are shown in figure 2.4.

In the example shown in this figure, at time t_1 , the pulses applied by the amplifiers to all three decades of the counter (the 100, 10 and 1 decades) are above the threshold level of the counters. As a result the counter registers $100 + 10 + 1 = 111$, and consequently the DAC produces an output voltage of 111 steps.

At t_2 again 111 steps are added but at t_3 the amplitude of the square wave is no longer sufficient to produce a pulse exceeding the threshold voltage of the 100 decade, so only the 10 and 1 decade are activated and 11 steps are added. This continues until at t_6 only the 1 decade is activated; this decade is activated up to t_{12} . So in our example V_i is equal to

$$\begin{aligned}
 2 \times 111 &= 222 \\
 + 3 \times 11 &= 33 \\
 + 7 \times 1 &= 7 \\
 \hline
 262 \text{ steps}
 \end{aligned}$$

By proper choice of the reference voltage, each step can be made equal to e.g. 1 mV, so if the counter reading is presented on a display a direct reading of the input voltage in mV is obtained.

The measurement is completed when the best possible compensation has been obtained, i.e. when the "1" decade does not change any further after a predetermined number of chopper cycles (e.g. 5). The next measurement is started by resetting the counters to zero.

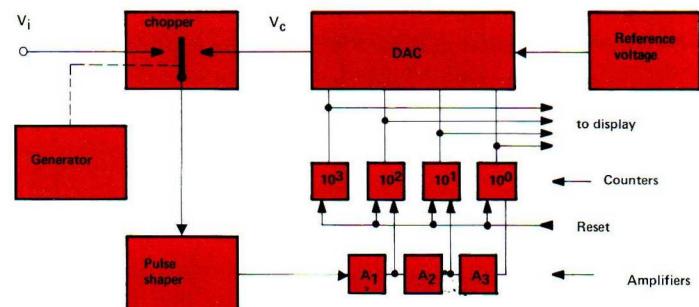


Fig. 2.3. Block diagram of a successive-approximation ADC

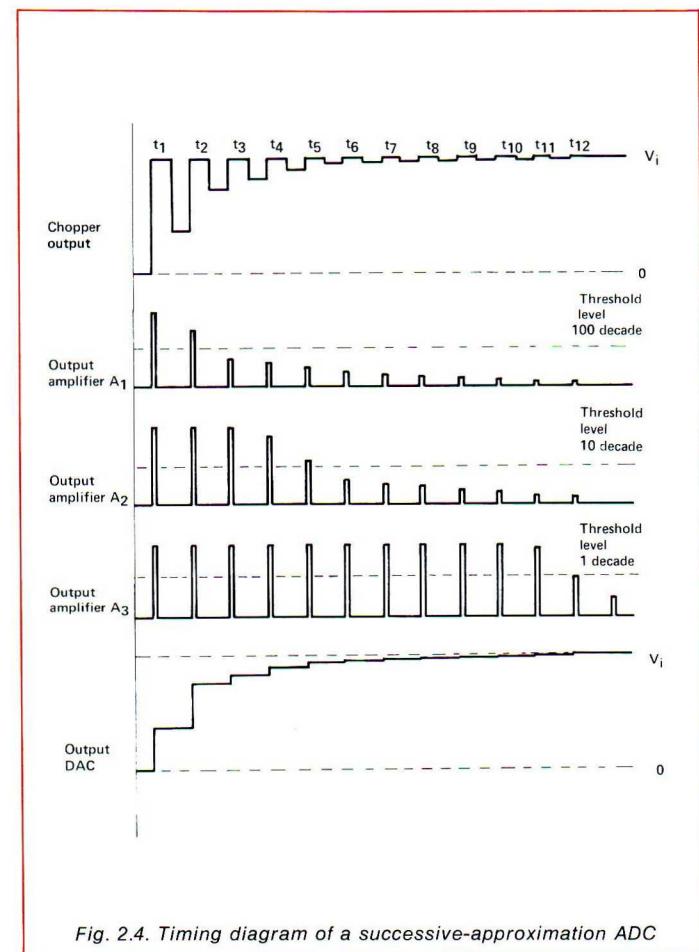


Fig. 2.4. Timing diagram of a successive-approximation ADC

Advantages and disadvantages

The features of this ADC do not differ much from those of the staircase-ramp ADC.

The main difference is that, thanks to the higher steps at the start of the measurement, this system can operate faster. This presupposes the use of a solid-state chopper to permit high-speed switching.

Remark

As with nearly all types of ADC, the successive-approximation ADC exists in several versions.

In some of these systems the input is compared with the output of a binary scaler connected to the reference voltage.

During conversion all bits are compared with the input voltage one at a time, starting with the most significant one: the input comparator then supplies an output indicating whether the input amplitude is greater or less than the output of the digital-to-analog converter. If V_i is less the bit is rejected, if V_i the greater the bit is accepted and stored. To illustrate this, we shall show how a voltage $V_i = 0.727 V_{ref}$ is measured, ($V_{ref} = 1 V$).

| Bit | Equal to | | Accepted | Sum |
|------------------|------------|-----------|----------|---------|
| Most significant | 0.5 | V_{ref} | yes | 0.5 |
| 2 nd | +0.25 | V_{ref} | no | 0.5 |
| 3 rd | +0.125 | V_{ref} | yes | 0.625 |
| 4 th | +0.0625 | V_{ref} | yes | 0.6875 |
| 5 th | +0.03125 | V_{ref} | yes | 0.71875 |
| 6 th | +0.015625 | V_{ref} | no | 0.71875 |
| 7 th | +0.0078125 | V_{ref} | yes | 0.72656 |

An additional advantage of this version is the constant conversion time (as all bits have to be scanned in all cases) and the immediate availability of a serial binary-coded output.

Variable-frequency method

This analog-to-digital converter makes use of a voltage-to-frequency conversion, the frequency being measured with a normal counter, figure 2.5.

The conversion is realized by integrating the input voltage with an operational amplifier (see chapter 1).

As soon as a fixed switching level V_s is reached the level detector passes a command pulse to a switch in parallel with capacitor C , causing the latter to be discharged; the switch then opens again and the cycle is repeated. Each time the switch is closed, the level detector feeds a pulse to the counter.

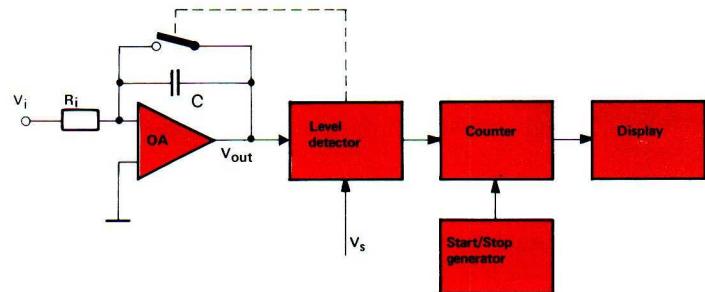


Fig. 2.5. Block diagram of a variable-frequency ADC

As $V_{out} = \frac{1}{R_i C} \int V_i dt$, (see chapter 1) the time t required to build V_{out} up to the constant switching level V_s is related to V_i as $t = \alpha \frac{1}{V_i}$ so the frequency f at which the

switch is closed is $f = \frac{1}{t} = \frac{V_i}{\alpha}$

By making the right choice of R_i , C and the switching level V_s , we can give α such a value that the output is suitable for direct reading of the input voltage.

To measure the frequency the counter unit must be started and stopped by signals supplied by an internal generator, that are separated by an exactly known time interval T , figure 2.6.

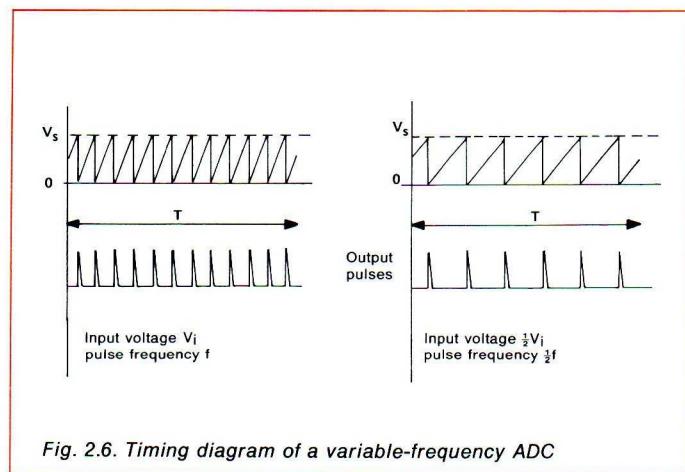


Fig. 2.6. Timing diagram of a variable-frequency ADC

Advantages and disadvantages

Since measurement is based on an integrating system the average AC component of the input tends to zero so the series-mode rejection can be high. Another positive point is that the pulse output of the level detector can easily be transferred through a guard without any need for DC coupling; this system is thus very suitable for use in instruments with guarded inputs. From a point of view of accuracy the system is not so attractive, as too many factors can influence the accuracy and the stability; e.g. R_i , C , V_s and the time T during which frequency is measured are of importance for the accuracy.

The high kick-back voltage at the input (due to the fast discharge of the integrating capacitor) can also be an objection.

Dual-slope integration

A widely used type of ADC is the dual-slope-integration ADC, based on integration of the input voltage for a constant time followed by measurement of the time required to discharge the integrating capacitor with a constant current. The block diagram of this set-up is shown in figure 2.7.

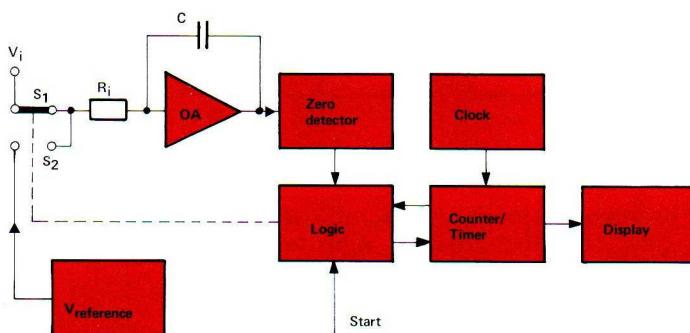


Fig. 2.7. Block diagram of a dual-slope-integration ADC

At the start of a measuring period the logic circuit closes switch S_1 , and at the same time starts the timer counting a fixed number of clock pulses corresponding to a time T . During time T with V_i connected to the operational amplifier the input voltage will be integrated in capacitor C , so at the end of this period V_c is directly proportional to V_i . At the end this period the logic now opens S_1 and closes S_2 so that the reference voltage is applied to the integrator. If we assume that V_{ref} is of opposite polarity to V_i , it follows that C starts to discharge. As the reference voltage is constant the discharge slope is constant too, figure 2.8.

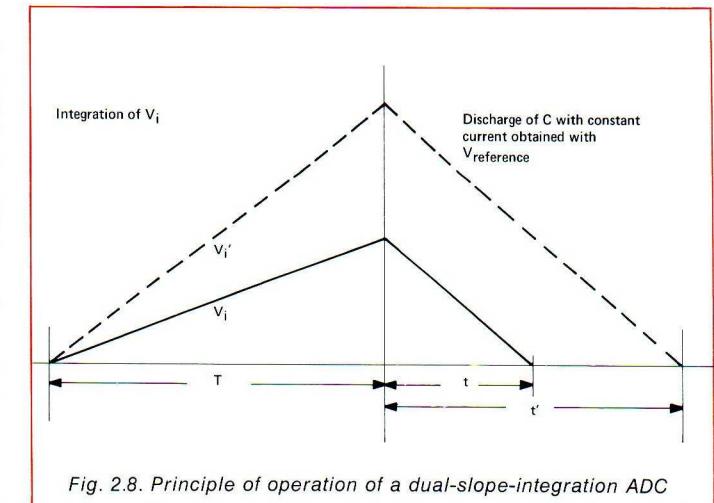


Fig. 2.8. Principle of operation of a dual-slope-integration ADC

The time t needed to discharge the capacitor completely is therefore a linear function of the value of V_c at the end of period T and hence of V_i .

We can describe this process by the following equations:

$$V_c \text{ (at the end of period } T) = \frac{1}{R_i C} \int_0^T V_i dt = \frac{1}{R_i C} V_i' T$$

(V_i' is the average value of V_i during T)

During discharge time t

$$V_c = \frac{1}{R_i C} \int_0^t V_{ref} dt = \frac{1}{R_i C} V_{ref} t$$

or $V_i' T = V_{ref} t$

$$\frac{V_i'}{V_{ref}} = \frac{t}{T}$$

The time t is measured by the counter, which is started immediately after time T is ended and stopped by the logic when the zero detector indicates that the capacitor is discharged.

When both t and T are measured by counting pulses from the same clock the only factor influencing the accuracy is in fact the reference voltage.

Advantages

The system has excellent series-mode rejection thanks to the integration over a longer period. This, together with the fact that the reference voltage is the only source of error, makes the system suitable for very accurate measurements.

Delta-pulse modulation-system

A system which is gaining in importance is delta-pulse modulation. It can be regarded as a mixture of several of the above-mentioned principles, and has many attractive features.

The block diagram is shown in figure 2.9.

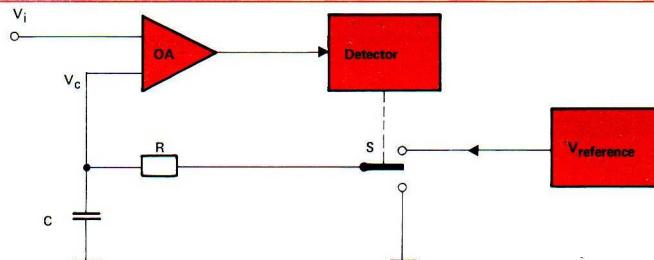


Fig. 2.9. Simplified block diagram of a delta-pulse-modulation ADC

An operational amplifier measures the difference between the input voltage V_i and the voltage across capacitor V_c . A polarity detector connected to the operational amplifier controls selector switch S. As long as V_i is more positive than V_c , switch S is in the upper position and the positive reference voltage is connected to C. When V_i is negative compared to V_c , C is connected to ground as S is in the lower position. This is a continuous process and as the gain of the operational amplifier is high V_c will always be nearly equal to V_i ; in fact the maximum difference will only be a few microvolt once the steady state has been reached.

If V_i is constant, the charge on capacitor C must be constant too (otherwise V_c would change).

In other words, the incoming charge Q_{in} must be equal to the discharge Q_{out} when measured over a certain period T:

$$Q_{in} = Q_{out}$$

Suppose that during this period T switch S is in the upper position for a time t and consequently in the lower position for a time $T-t$; we may thus write:

$$Q_{in} = \frac{V_{ref} - V_c}{R} t \quad \text{and}$$

$$Q_{out} = \frac{V_c}{R} (T-t)$$

and since $Q_{in} = Q_{out}$,

$$(V_{ref} - V_c) t = V_c (T-t) \quad \text{or}$$

$$V_i = V_c = \frac{t}{T} V_{ref}$$

So if we can measure the ratio t/T we also know V_i (since V_{ref} is known).

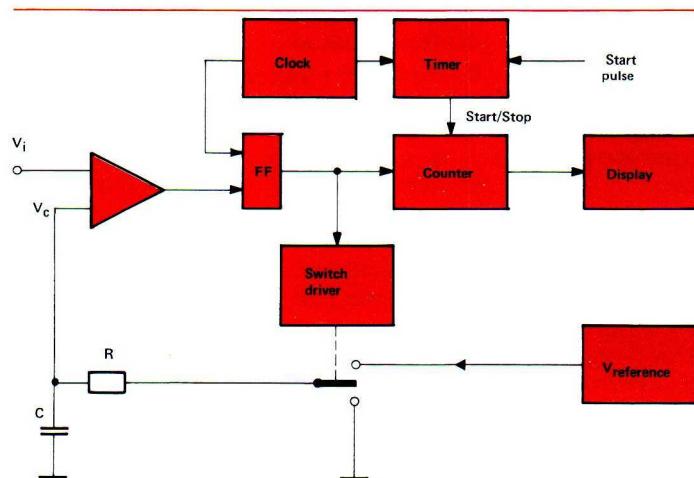


Fig. 2.10. More detailed block diagram of a delta-pulse-modulation ADC

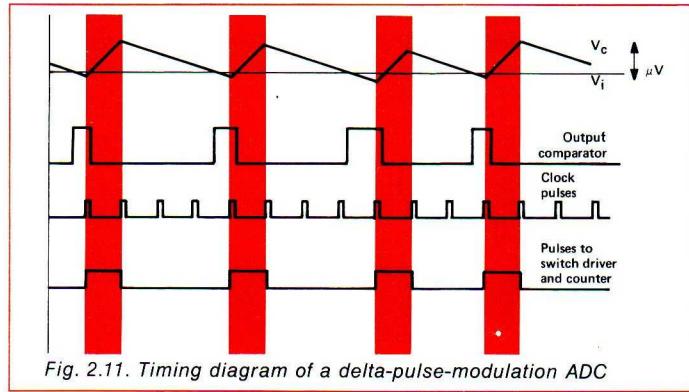


Fig. 2.11. Timing diagram of a delta-pulse-modulation ADC

The measurement of this ratio is reduced to a simple digital counting problem with the aid of the circuit of figure 2.10.

The detector is now a flip-flop that can be set as long as $V_i > V_c$ and not when $V_i < V_c$ (resetting is always possible). This flip-flop operates as a binary scaler and so produces pulses with a duration equal to the interval between two clock pulses (as long as $V_i > V_c$).

These pulses control the switch driver, which brings the switch to the upper position so that C is charged from the reference source during these pulses. During the rest of the time, C is discharged.

After an external command, the timer starts to count a fixed number, e.g. 4000, of clock pulses. At the same time the counter is set for counting the output pulses of the flip-flop. When the timer has reached the end of its count, the counter is stopped again. If the reading of the counter is n, this means that the capacitor has been charged during n out of the 4000 clock intervals

Hence, $t = n\tau$ and $T = 4000 \tau$

$$\text{so } \frac{V_i}{V_{\text{ref}}} = \frac{t}{T} = \frac{n}{4000} \text{ or } V_i = \frac{n}{4000} V_{\text{ref}}$$

As the maximum value of n in our example is 2000 (which means that all pulses are passed on to the switch driver), V_{ref} should be chosen as twice the maximum value of V_i that can be measured on the scale used at that moment. Figure 2.11. illustrates the operation of this system when V_i is about $\frac{1}{4}$ of V_{ref} .

Note

The linear ramps of the sawtooth voltage in fig. 2.11 and the linear relations in the above calculations are due to the fact that near the equilibrium of V_c and V_i the changes in voltage across C are of the order of microvolts (high gain of op. amp.). Whereas V_c and V_{ref} themselves may amount to some volts.

Advantages

This system has the same advantages as the dual-slope principle: since it is an integrating system, it has a good series-mode rejection; and its accuracy depends only on the reference voltage. In addition, we may mention that the design of this circuit lends itself very well to the use of modern technology such as large-scale integrated circuits (LSI circuits).

Over-all comparison

The pro's and con's of the different systems are summarized in the following table:

| System | Accuracy | Input impedance | Series-mode rejection (SMR) | Kick-back signals | Speed |
|---|----------|-------------------------------|-----------------------------|-------------------|--------|
| Staircase ramp (compensation) | high | (not constant) low to high | low | medium | medium |
| Successive approximation (compensation) | high | (not constant) low to high | low | low | high |
| Variable frequency | medium | limited | high | high | medium |
| Dual-slope integration | high | very high | high | low | medium |
| Delta-pulse modulation | high | high | high | low | medium |

Questions

Check the answer(s) you believe to be correct.

Q.2.1. The accuracy of a staircase-ramp ADC is high because

- A it measures the instantaneous value of the input voltage
- B the DAC and the reference voltage source are the only sources of error
- C the number of steps in the staircase ramp can be made very high

| | |
|---|--|
| A | |
| B | |
| C | |

Q.2.2. The main advantage of a successive-approximation ADC is

- A it is independent of its chopper frequency
- B it has a high series-mode rejection
- C it can operate faster than other ADC's

| | |
|---|--|
| A | |
| B | |
| C | |

Q.2.3. Which of the following statements concerning the variable-frequency ADC are true?

- A the series-mode rejection can be high
- B the kick-back signals are high
- C the accuracy is only dependent on that of the reference voltage

| | |
|---|---------------|
| A | true/not true |
| B | true/not true |
| C | true/not true |

Q.2.4. The dual-slope ADC is widely applied because

- A its accuracy depends only on that of the reference voltage
- B it is fast
- C it has a high series-mode rejection

| | |
|---|--|
| A | |
| B | |
| C | |

Q.2.5. The delta-pulse modulation ADC has the advantage(s) that:

- A it gives a high read-out resolution
- B it has a high series-mode rejection
- C its accuracy depends only on that of the reference voltage

| | |
|---|--|
| A | |
| B | |
| C | |

The answers to the problems will be found at the end of this book.

PM 2524 automatic digital V A Ω meter

MADE IN HOLLAND

PHILIPS

2429

mV
V
k Ω
M Ω
mA
DC

HOLD
RANGING
AUTO UP DOWN

POWER

ON
OFF

0

V Ω

A

DC

MAX
1000V 500V

V = V \sim

mA = mA \sim

0

Ω DC



Chapter 3

Automation in voltmeters

One of the advantages of digital multimeters is their ease of operation. The reading is easy to take and does not lend itself to errors of interpretations. Moreover, the number of ranges is limited because of the ranging in steps of 10 (instead of the $\sqrt{10}$ steps used for analog instruments). Demand from users and advances in digital circuitry for simple forms of computation signalling and control led to a further development, in which more and more automatic functions were incorporated in digital voltmeters.

Nearly all instruments today have automatic polarity display and automatic decimal point positioning, while many have auto ranging and auto zeroing too. In this chapter we will briefly describe the principles underlying these various automatic functions.

Automatic polarity indication

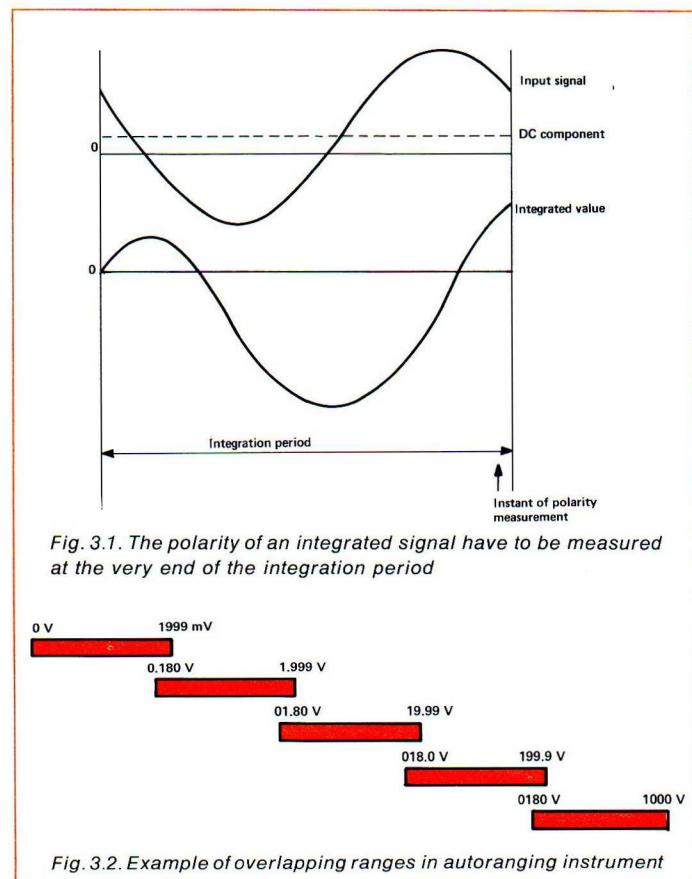
The polarity indication is generally obtained from the information in the ADC (analog-to-digital converter). For integrating ADC's only the polarity of the integrated signal is of importance. The polarity should thus be measured at the very end of the integration period (see fig. 3.1.). As the length of the integration period is determined by counting a number of clock pulses, it is logical to use the last count or some of the last counts to start the polarity measurement. The output of the integrator is then used to set the polarity flip-flop, the output of which is stored in a memory until the next measurement is made.

Autoranging

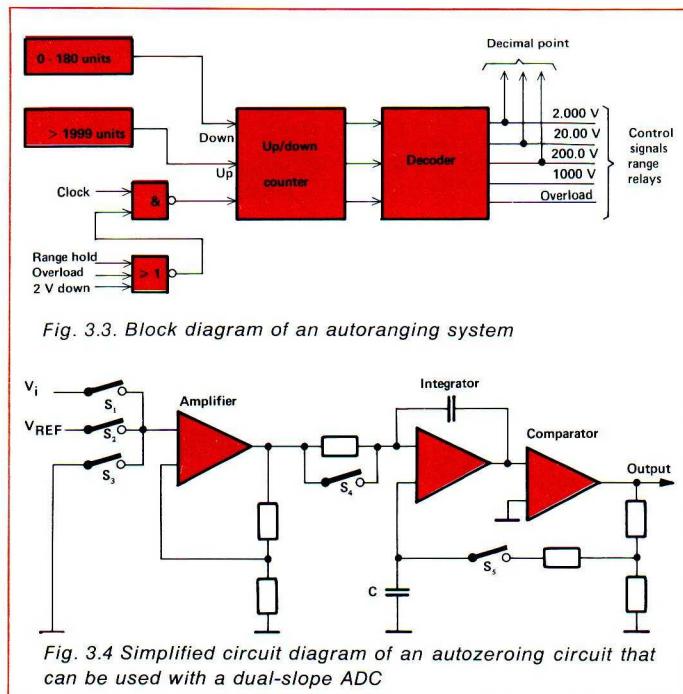
The object of autoranging is to get a reading with optimum resolution under all circumstances, e.g.: 170 mV should be displayed as 170.0 and not as 0170.

Let us take the example of a $3\frac{1}{2}$ digit display, i.e. one with a maximum reading of 1999. This maximum means that any higher value must be reduced by a factor of ten before it can be displayed, e.g. 201 mV as 0201. On the other hand, any value under 0200 can be displayed with one decade more resolution e.g. 195 mV as 195.0. In other words, if the display does not reach a value of 0200 the instrument should automatically be switched to a more sensitive range, and if a value of higher than 1999 is offered next less sensitive range must be selected.

In general practice the lower limit is taken lower than 0200, e.g. 0180. Otherwise, a voltage exhibiting slight fluctuations around 200 would be displayed successively as e.g. 199.9, 0200 and 0201, which would be confusing. By introducing an overlap in the ranges (see fig. 3.2.), we ensure that all values are displayed in the same range, in the above example as 0199, 0200 or 0201. Values around 180 also give a stable display e.g. 1798, 1800 and 1807.



The design of an automatic ranging system is indicated in the block diagram of fig. 3.3. The information contained in the counter of the ADC yields a control pulse for "down" ranging when the count is less than 180 and one for "up" ranging when the count exceeds 1999 units. The up/down counter of the autoranging circuit reacts to this information at the moment that a clock pulse (a pulse at the end of the measuring period, also used to transfer new data to the memory) is applied, and the new information is used to set the range relays via the decoder. At the same time the decimal point in the display is adapted to the new range. When more than one range step has to be made, several measuring periods are needed to reach the final result.



Clock pulses, and so autoranging can be inhibited e.g. by a manual "range hold" command, by a signal that exceeds the maximum range (only for "up" counts), and of course by reaching the most sensitive range – but then only for "down" counts.

Autozeroing

Each user of a voltmeter expects the instrument to indicate zero when the input is short-circuited. In a digital voltmeter with a maximum reading of 1999, a zero error of 0.05% of f.s.d. is sufficient to give a reading of 0001.

For this reason, and in the interests of optimum accuracy with low-valued readings, a zero adjustment is necessary. To increase the ease of operation, many instruments contain an automatic zeroing circuit nowadays. In a system used in several multimeters, the zero error is measured just before the real measurement, and stored as an analog signal e.g. in the integrator. A simplified circuit diagram of a circuit that can be used for this purpose is given in fig. 3.4. for a dual-slope ADC. Before the real measurement is made, switches S_3 , S_4 and S_5 are closed for e.g. 50 ms, thus earthing the input, giving the integrator a short RC time and connecting the output of the comparator to capacitor C. This capacitor is now charged by the offset voltages of the amplifier, the integrator and the comparator; and when switches S_3 , S_4 and S_5 are opened again to start the real measurement, the total offset voltage of the circuit (=zero error) is stored in this capacitor, so that the real input voltage is measured correctly.

The fully automatic instrument

A multimeter with automatic polarity indication, automatic zero correction and automatic ranging (of course coupled with automatic decimal-point indication) only needs a signal applied to its input and a command as to what quantity (e.g. V_{DC} , V_{AC} , I or R) to measure; it does all the rest itself. All these features can already be found in low cost instruments such as the Philips PM 2514, an auto-ranging multimeter which also has facilities for temperature measurement.

The organization of the digital part of a voltmeter

We have seen above how an analog signal can be converted into a digital one, and how some of the functions of the instrument can be automated with the aid of the information contained in ADC. We will now study how the digital part of a typical instrument is organized so as to produce a display or a digital output signal, (see fig. 3.5.). Before a measurement can begin, the functions of the instrument must be set; i.e. we must select the quantity to be measured (e.g. voltage), the ranging mode (automatic or manual) and the start mode (internal or with an external trigger signal). This can be done by front-panel controls, or via a remote-control input. In both cases, the signals are fed to the "function control" unit, while the information on ranging is passed to the "range control" unit.

Let us assume that in the instrument in question, the ADC is of the dual slope integration type, and that a choice can be made between the following combinations:

| Measurements/s | Integration time | Clock frequency |
|----------------|------------------|-----------------|
| 4 | 100 ms | 200 kHz |
| 20 | 20 ms | 1 MHz |
| 200 | 2 ms | 1 MHz |

It will be clear that as the number of measurements per second increases the integration time must be reduced and that it is useful to increase the clock frequency at the same time to maintain good resolution. To select the desired combination, information on the number of measurements per second must be fed to the start oscillator and the clock oscillator.

The latter constantly supplies clock pulses to the programming unit, while the former is also free-running when the DVM is set for internal start but waits for an external trigger signal when the DVM is set for external control. Let us now follow (with reference to fig. 3.5.) the various steps involved in the performance of a measurement, for the case that the instrument is set for autoranging and external triggering.

An incoming trigger pulse causes the start oscillator to

deliver a pulse to the programming unit, and a measurement is started. The programming unit starts both the counter and the ADC. The ADC is connected to the input. The counter counts the clock pulses to determine the integration time and sends two signals back to the programming unit, one just before the end of the integration period, and the other at the end of this period. The first signal is used by the programming unit to activate the polarity detector, which determines the polarity of the integrated signal, while the second serves to switch the ADC input from input signal to reference signal. At the same time, the counter is reset to zero and starts counting the down integration time of the ADC until it is stopped by the zero detector signal of the ADC (see under dual-slope integration in chapter 2). At that moment, the programming unit compares the counter reading with the autoranging limits, and passes an up or down signal to the range control if necessary. This unit switches the input range switches via a relay and triggers the start oscillator for a new measurement in a more sensitive or less sensitive range. In the meantime, the programming unit will also have reset the counter.

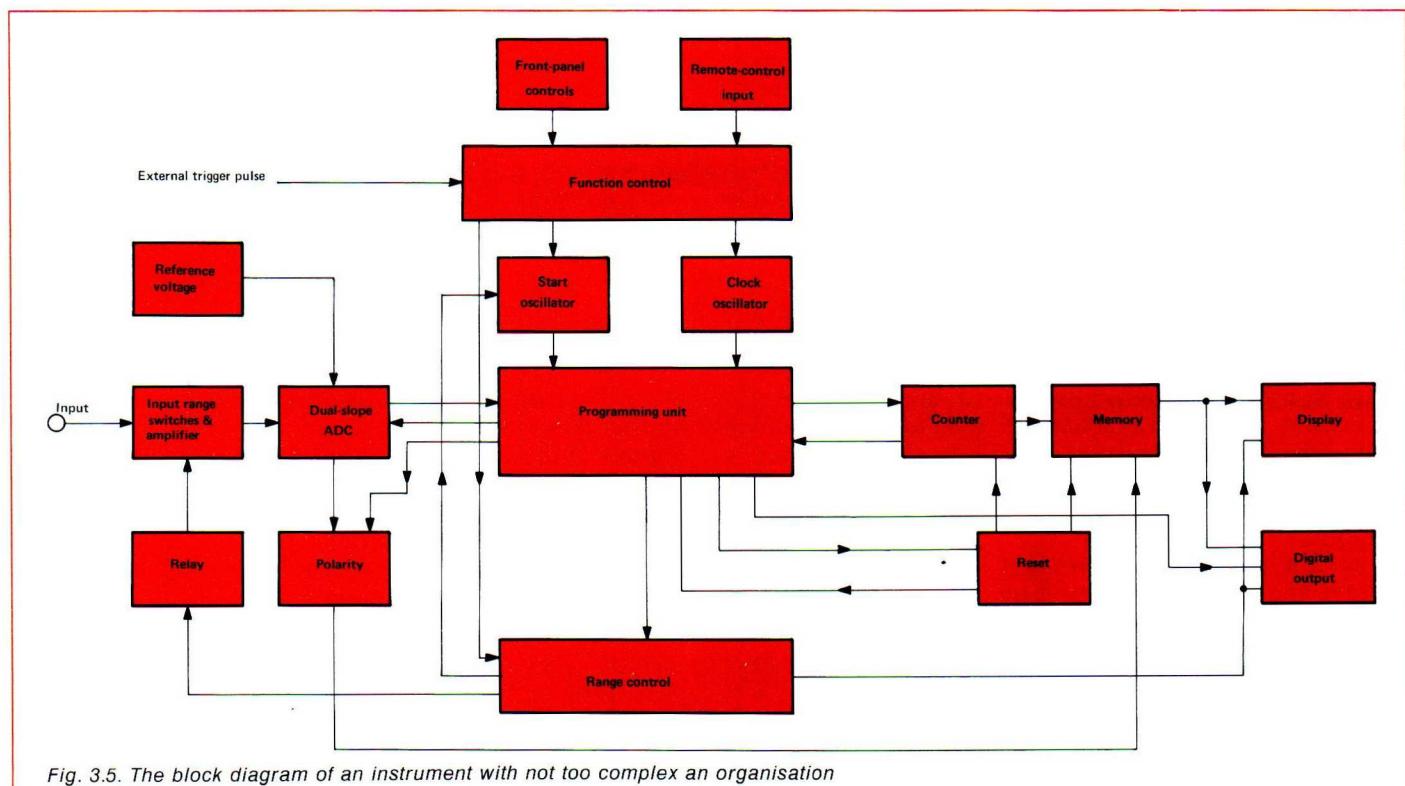
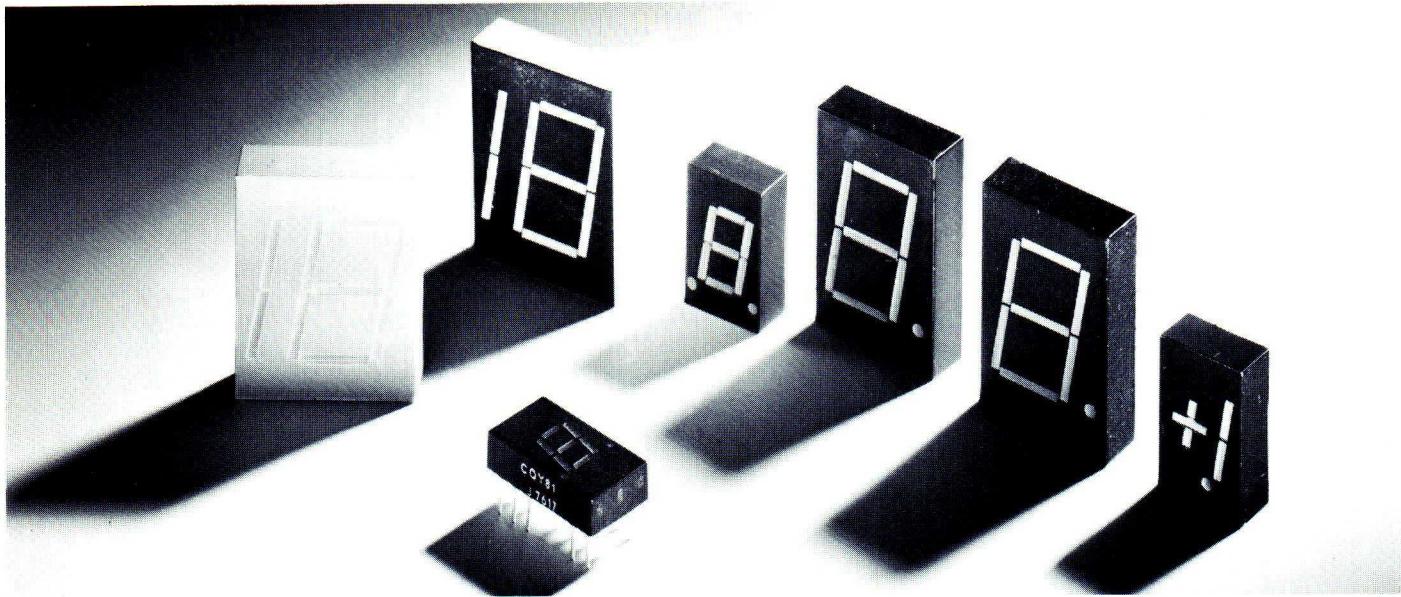


Fig. 3.5. The block diagram of an instrument with not too complex an organisation



This process continues until a measurement has been made the result of which is within the autoranging limits. The programming unit then transfers the new data from the counter to the memory, together with the polarity information, so as to make them available to the display unit and the digital output. Finally, the programming unit delivers a transfer pulse to the digital output, to "warn" an instrument connected to this output (e.g. a printer) that new data have been made available.

Display devices

The value measured by a digital voltmeter is generally displayed in the form of a + or - sign followed by a number of digits with the decimal point at the appropriate place in between. There are several devices which can be used to realize such a display, but the most common ones are:

- Gas-filled decade tubes,
- 7-segment light-emitting-diode (LED) displays, and
- Liquid-crystal displays (LX or LCD).

Gas-filled tubes have been used right from the start in DVM's, and are still in use because of the nice shape and high brightness of the figures they display. Each of the ten digit-shaped electrodes in these tubes is connected as a separate cathode and all have one common anode. By applying a voltage to the anode and grounding the appropriate one of the cathodes, we cause the corresponding digit to light up. In other words, we need a one-out-of-ten code for driving this type of display.

The disadvantages of this type of display are the high voltage needed (which is a particularly big drawback in fully transistorized voltmeters), and the rather high power consumption. Moreover the reading angle (i.e. the maximum angle between the viewing direction and a perpendicular line to the display) is limited, as the display digits are not coplanar but arranged one behind the other. Light-emitting-diode (LED) displays and later liquid-crystal (LX) displays were developed to solve these problems.

In both types of displays, a digit is built up from a maximum of 7 segments, see fig. 3.6.

They thus need an n-out-of-7 code to drive them.

The main advantages of an LED display are that all digits are in one plane, and that it is an active light source available in different colours.

Thanks to their physical properties, LX displays have a very low power consumption; they are thus highly suitable for battery-operated equipment. A drawback is that they are not active light sources being dependent on ambient light or an internal light source, for operation in the reflective or transmissive mode.



Fig. 3.6. How digits are built up from the seven segments of an LED or LX display

Both LED and LX displays can be driven by low voltages; they can thus be connected directly to transistor or IC circuitry.

This is all the more important since all types of display need some form of decoding (from BCD to 1 out of 10 or n out of 7).

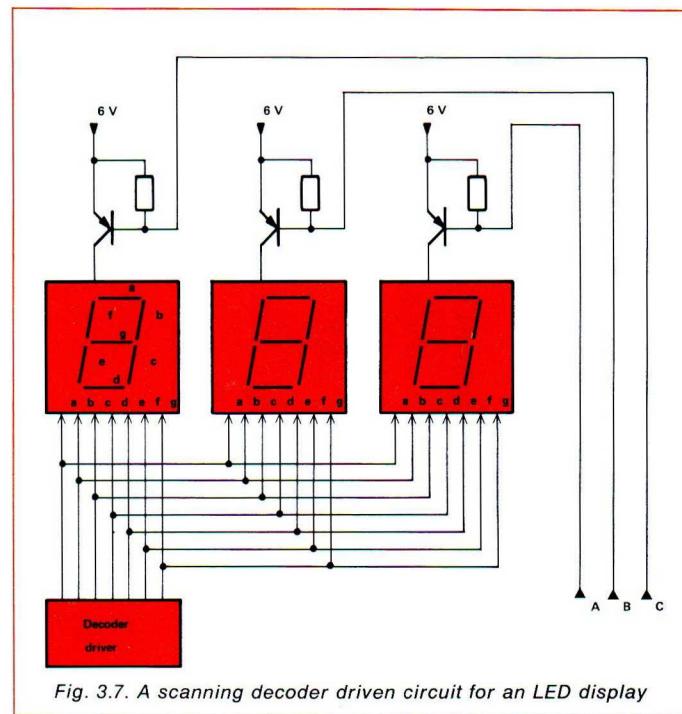


Fig. 3.7. A scanning decoder driven circuit for an LED display

The decoder

The decoders and the display driver are generally combined in present-day voltmeter circuitry.

The driver function is only required to adapt the output of the decoder to the voltages and currents needed by the display. The decoder function ensures transformation of the binary output of the counter or memory into a code suitable for the display. This can be realized by simple diode or AND-gate matrices. An example of a binary-to-decimal decoder has been described in Part 1, Chapter 8 of this course.

Special integrated circuits now contain full decoding and driving functions.

However, even with these IC's a large number of discrete component and even more interconnections are needed to drive e.g. a $4\frac{1}{2}$ -digit 7-segment display. A scanning mode of operation is sometimes used to reduce this large number of connections.

This means that one single decoder-driver is connected to all display units in parallel. An individual anode switch for each display unit connects the latter in series with the supply voltage, synchronously with the appropriate coded data (see fig. 3.7.).

If e.g. the number 347 has to be displayed, the digit 3 in BCD code is first applied to the decoder and control line A is activated; then 4 is applied to the decoder and control line B is activated; and finally 7 is applied to the decoder and control line C is activated. The use of this serial-mode drive is not apparent from the display as long as the repetition frequency is high enough. The brightness of the display is lower than normal, of course, as the individual digits are only energized during part of the cycle.

Questions

Check the answer(s) you believe to be correct.

Q.3.1. In an autoranging DVM the ranges overlap in order to ensure:

- A a brighter display
- B a higher resolution
- C a more stable display at the ends of the ranges

| | |
|---|--|
| A | |
| B | |
| C | |

Q.3.2. Autozeroing is used in a DVM because:

- A it gives more accurate results
- B it gives a more stable display
- C it compensates for internal offset errors

| | |
|---|--|
| A | |
| B | |
| C | |

Q.3.3. When the number of measurements per second is increased in a programmable DVM,

- A the accuracy increases accordingly
- B it is useful to increase the internal clock frequency to
- C the integration time of the ADC must be decreased

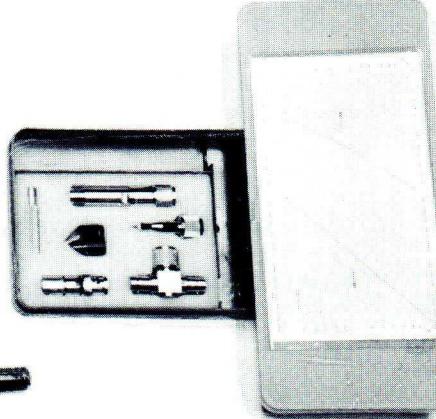
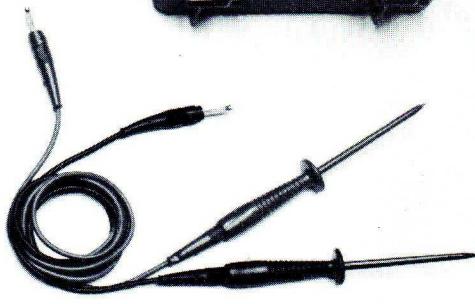
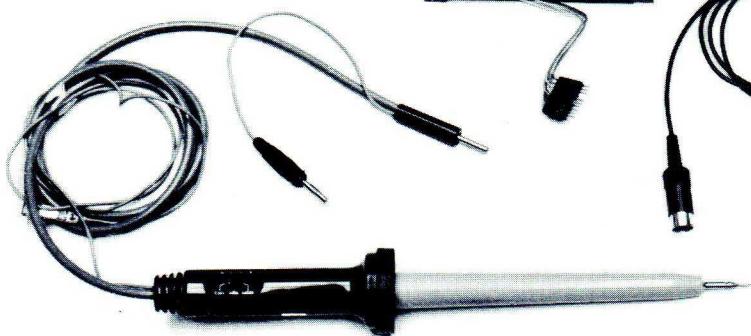
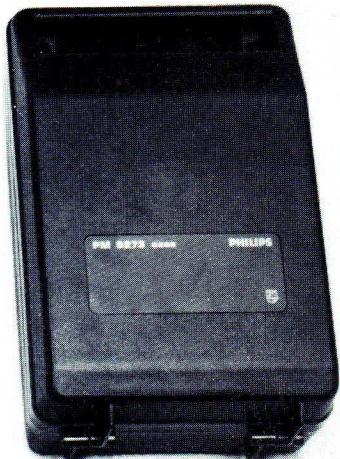
| | |
|---|--|
| A | |
| B | |
| C | |

Q.3.4. LED or LX display units are replacing gas-filled decade tubes ("nixies") because:

- A LED displays have a more elegant shape
- B nixies have a higher power consumption
- C nixies require a more complicated driving circuit

| | |
|---|--|
| A | |
| B | |
| C | |

The answer to the problems will be found at the end of the book.



Chapter 4

Digital multimeter circuits

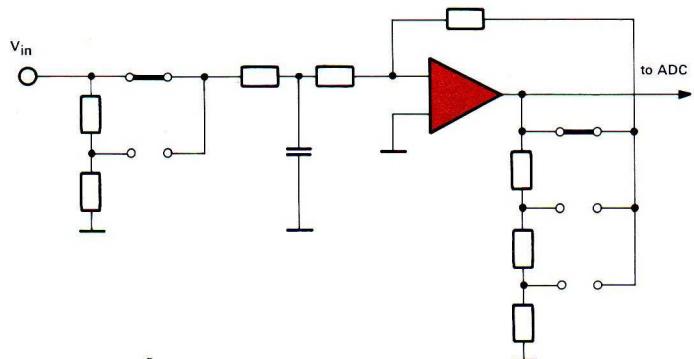


Fig. 4.2. An attenuator with an operational amplifier connected as a variable gain pre-amplifier.

As we have seen, an ADC is only capable of handling DC voltage inputs.

When, as in a multimeter, other quantities such as DC current, AC voltage and current and resistance have to be measured, these quantities first have to undergo analog conversion to a DC voltage.

Although the subject of analog conversions might be thought to be a bit outside the scope of a digital instrument course, we need to say a little about these techniques in order to provide a better basis for understanding the considerations concerning the accuracy of digital voltmeters given in chapter 5.

DC Voltage attenuator

The DC signals that can be handled by the input of an ADC are in general limited to less than 10 V. This means that DC voltages of more than 10 V have to be attenuated in the input stage of a DMM before they can be passed on to the ADC. A simple form of attenuator which can be used for this purpose is shown in fig. 4.1.

In the 200 mV and 2 V settings, the input signal is applied directly to the ADC. In the other settings, it is attenuated first. It will be seen that the attenuator offers a fixed input impedance of 10 M Ω in all settings.

A voltage overload device is included in the circuit to protect it against excessive input voltages.

A second solution which is often used is to employ an operational amplifier connected as a pre-amplifier with adjustable gain, as described in chapter 1. The number of attenuation steps that can be realized in this way is limited by the dynamic range of the amplifiers; a voltage divider giving a one-step preattenuation of e.g. 1000:1 must therefore be included in the circuit to allow the full range to be covered. An example of such a circuit is shown in fig. 4.2.

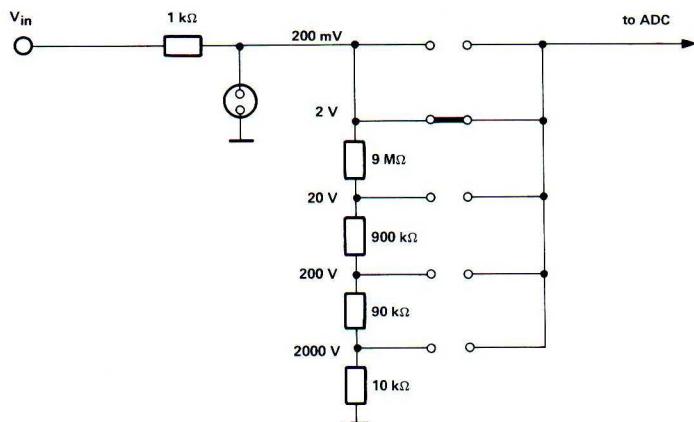


Fig. 4.1. A simple DC attenuator.

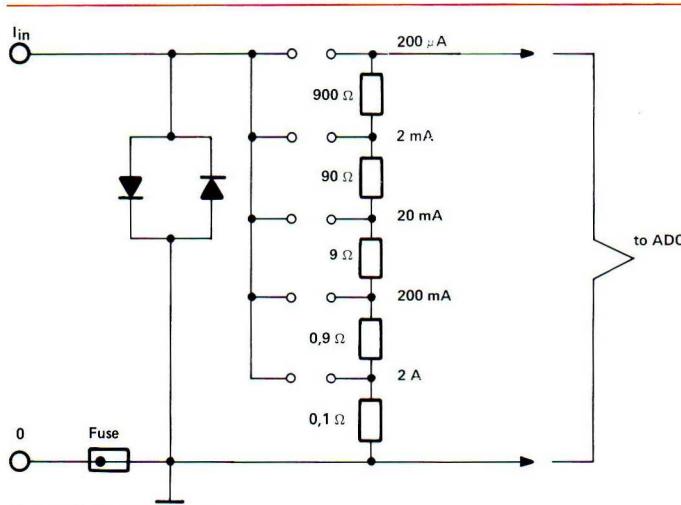


Fig. 4.3. Current-to-voltage converter.

Current-to-voltage converter

DC currents are converted into voltages by shunts in such a way that the voltages across the shunt at full range is the same for all current ranges, and is as low as possible. To provide overload protection in the lower current ranges, two diodes in anti-parallel are added to the circuit: before the voltage across the shunt resistors can assume a dangerous value, one of the diodes starts conducting and the fuse will blow (fig. 4.3.).

AC-DC converter

A widely used system for AC-DC conversion is the average detector described below. More reliable results are given by a true RMS converter, as the accuracy of the latter is independent of the waveform (see chapter 7); but this solution is more expensive. In the average detector, two diodes D_1 and D_2 are arranged in the feedback loop of an operational amplifier OA (fig. 4.4.).

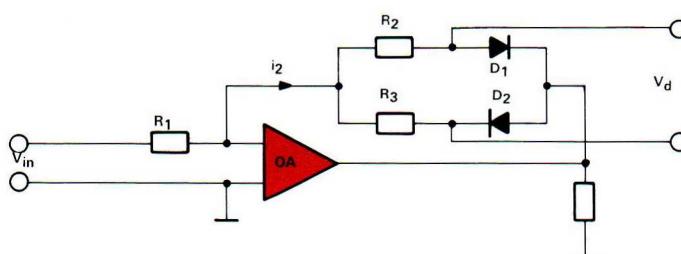


Fig. 4.4. An average detector as AC-to-DC converter.

$$i_2 = \frac{V_{in}}{R_1} \text{ (see chapter 1)}$$

i_2 passes through R_2 or R_3 depending on the polarity of the input signal, so:

$$V_d = i_2 \times R_2 + 0 \times R_3 \text{ or } V_d = 0 \times R_2 + i_2 \cdot R_3;$$

$$\text{and when } R_2 \text{ and } R_3 \text{ are equal, then } V_d = V_{in} \frac{R_2}{R_1}$$

V_d is thus a linear function of V_{in} .

The advantage of this system is that it gives linear detection from very low AC levels. This can be shown as follows.

As long as the diodes are cut off, the resistance in the feedback loop of the amplifier is very high, so the gain of the circuit is approximately equal to the open-loop gain and is thus very high too (of the order of 5×10^7). The diodes start to conduct at about 0.5 V, i.e. at an input

$$\text{voltage } V_{in} \text{ of } \frac{0.5 \text{ V}}{\text{open-loop gain}} \approx 10 \text{ nV};$$

voltages of or above this value can thus be detected.

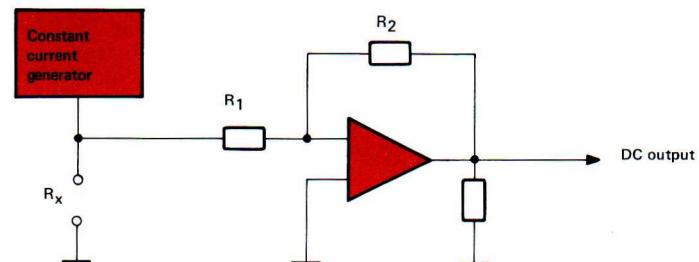


Fig. 4.5. Resistance-to-voltage converter.

Resistance-to-voltage converter

Resistances are measured by passing a constant current through the unknown resistance R_x and measuring the voltage across it. A possible circuit is given in fig. 4.5. As very high resistance ranges would require very low currents with this set-up, an alternative version can be used making use of the fact that the gain of the operational amplifier depends on the ratio R_2/R_1 (see fig. 4.6.). Now the input voltage is constant (fixed current through fixed resistance), while the output voltage is $R_x \frac{V_{in}}{R_1}$, i.e. proportional to R_x and the proportionality constant here is far lower than in fig. 4.5.

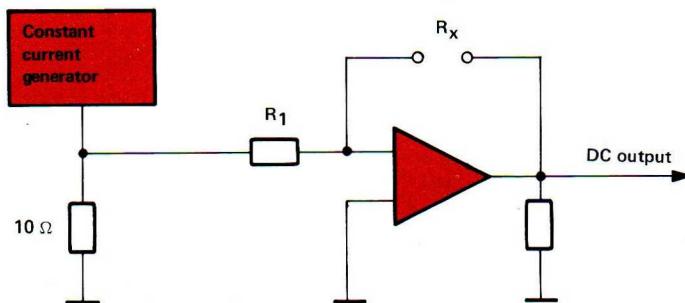


Fig. 4.6. Resistance-to-voltage converter for very high resistance ranges.

HF-to-LF converter

For the sake of completeness, we will close this chapter with a description of a system that can be used to permit HF measurements; such a system is used e.g. in the Philips Multimeter PM 2527.

The DMM circuitry which we have describing so far is not suitable for the measurement of very high-frequency voltages. However, this problem can be got round in an elegant way with the aid of the circuit of fig. 4.7.

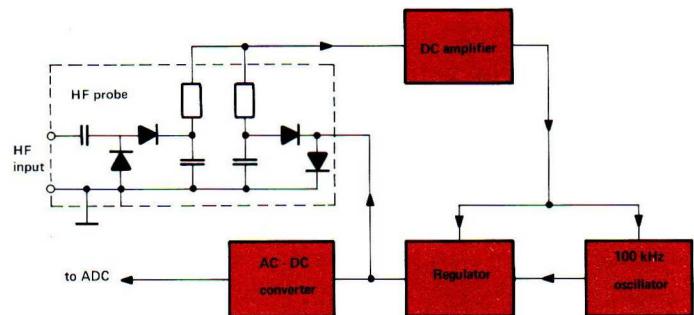


Fig. 4.7. HF-to-LF converter.

The input HF signal and an internally generated 100 kHz signal are compared after detection in the HF probe. The resulting difference signal is fed to the DC amplifier, whose output is used to control the amplitude of the 100 kHz signal. As the loop gain is high, the amplitude of the resulting 100 kHz signal is virtually the same as that of the original HF one. All we now have to do is to measure the amplitude of the LF signal in the normal way in the LF measuring circuit. Linear measurements are obtained from a level of some mV.

Questions:

Check the answer(s) you believe to be correct.

Q.4.1. Figure 4.8 shows a full-wave rectifier circuit. The diodes have a forward voltage drop of 0.5 V; they will therefore start to conduct at an input voltage V_{in} of:

- A 50 mV
- B 1 μ V
- C 5 μ V

| | |
|---|--|
| A | |
| B | |
| C | |

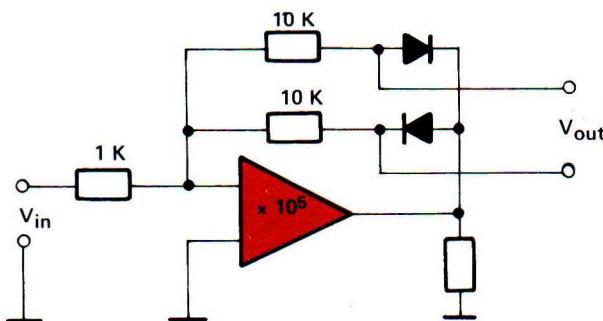


Fig. 4.8.

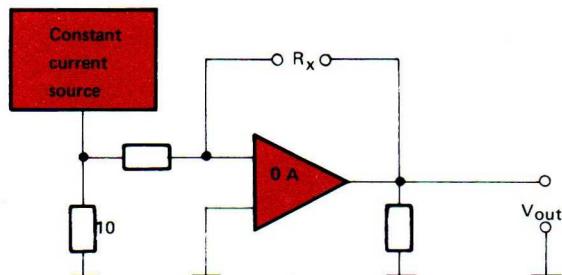


Fig. 4.9.

Q.4.2. The resistance-to-voltage converter of figure 4.9 is used to measure:

- A high resistances
- B low resistances
- C both A and B

| | |
|---|--|
| A | |
| B | |
| C | |

The answers to the problems will be found at the end of the book.



Chapter 5

Accuracy of digital voltmeters

Introduction

One of the main features of digital voltmeters is their high accuracy under optimum measuring conditions, ranging from 0.1% to 0.001% depending on technology and price class. Before discussing the various factors which can influence this accuracy, it is useful to consider how precisely the accuracy of such an instrument is specified. Let us take the example of a $4\frac{1}{2}$ digit DVM with a specified accuracy of:

$\pm 0.05\%$ of reading $\pm 0.02\%$ of range.

When we speak of a $4\frac{1}{2}$ digit instrument, we mean that the maximum reading is 19999. The $\pm 0.02\%$ of range in the above expression is thus equivalent to ± 4 in the last digit.



At the top end of the range, the maximum error will hence be $0.05\% \text{ of } 19999 + 4 = 14$ or 0.07%. At the lower end of the range, just before we switch to a more sensitive range, however, the maximum error is:

$0.05\% \text{ of } 2000 + 4 = 5$ or 0.25%.

In general, moreover, the accuracy is specified for certain standard operating conditions, of which the temperature is the most important.

A typical specification in this respect is:

- Reference temperature $23^\circ\text{C} \pm 1^\circ\text{C}$
- Temperature coefficient 0.005% of reading/ $^\circ\text{C}$

This means that if the instrument is used on a summer day with an ambient temperature of 34°C , an additional error of $(34 - 24) \times 0.005\% = 0.05\%$ can be expected.

The maximum over-all error in our example will then be 0.12% at the top end of the range and 0.30% at the lower end.

Another point included in the specification is often the long-term stability, i.e. the possible additional error due to aging effects. This could be specified as:

Long-term stability = 0.01% of reading over 90 days.

In this case, we add 0.01% to the over-all error if our instrument was last calibrated about three months ago. It is coming to be general practice to indicate a recalibration period instead of the long-term stability; the latter then has to be incorporated into the accuracy under standard conditions. In our example, the over-all accuracy would then be specified as:

Accuracy = $\pm 0.06\%$ of reading $\pm 0.02\%$ of range.

Recalibration period 90 days.

In the above examples, we calculated the worst-case error we can expect when we apply a voltage from a standard DC generator to the input of the voltmeter. However, it should be realized that, using a DC standard, measuring conditions are more or less ideal, since:

- the source impedance is low
- there is no common-mode or series-mode signal

In general, however, measurements are not made under such ideal conditions. We will now consider what effect it has on the behaviour of the instrument if the source impedance is not very low or series-mode or common-mode signals are present. The influence of the offset current, which is related to that of the source impedance, will also be discussed.

Sources of error in DC voltage measurements

Load error

The source impedance has the same effect on the accuracy of a digital voltmeter as on an analog one. If we want to make full use of the accuracy of digital equipment we should realize that a ratio of 1:1000 between the source impedance and the input impedance of the voltmeter will double the error of an 0.1% instrument. Hence, an input resistance of at least $10 \text{ M}\Omega$ is a must for a digital voltmeter.

Offset-current errors

Ideally, we would like the input of a voltmeter to be a true "one-way" device so that there is no feed back from the voltmeter into the measuring circuit. In practice, however, a small current does flow from the instrument in the measuring circuit we call this the offset current. The offset current is only an important source of error when measurements are made in high-ohmic circuits. Let us take the example of a 1 nA offset current passing through an impedance of 100 MΩ. The result is a voltage drop of 100 µV, or an error of 10 in the last digit of an instrument with 10 µV resolution see fig. 5.1. We should note that offset current of 1 nA is a small one. In some cases, the value will be far higher. Another problem is that the offset current is generally temperature-dependent. It is usually possible to compensate for this current at a given temperature, but compensation over the whole range of possible ambient temperature is very difficult.

It should be realised that a high input impedance in an electronic multimeter or electronic voltmeter is only useful if the offset current is very low. Otherwise, the influence of the offset current can be far higher than that of the load.

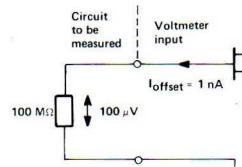


Fig. 5.1.

Series-mode errors

It often happens that the DC signal we want to measure also has an AC component, which is known as the "series mode" signal. How can we get rid of this series-mode error? One way would be to filter the AC component out; but this method has the disadvantage that the response to a change in the DC component is slow. The best solution for use in multimeters is the integrating analog-to-digital converter. By integrating the AC signal over a complete number of periods, we ensure that the resultant AC signal is zero.

The extent to which the AC signal (series-mode signal) is rejected by an instrument is expressed in terms of the series-mode rejection ratio (SMRR), which is generally somewhere in the range between 100 : 1 (40 dB) and 1000 : 1 (60 dB). As the SMRR is frequency-dependent, it is generally quoted at a specified frequency.

Series-mode are theoretically completely rejected by integrating DVM's at frequencies for which the integration time is equal to a multiple of the period. For example, series-mode signals of 50 or 60 Hz (with a period of 20 and 16 2/3 ms, respectively) are rejected very well with an integration period of 100 ms.

At intermediate frequencies, the SMRR is appreciably lower, but increases with increasing frequency. This is of vital importance, as at the low end of the frequency-scale the fundamentals and harmonics of mains frequency are the most important sources of error, while random noise has mainly higher-frequency components, so that both interference signals are well rejected by an integrating ADC. Fig. 5.2 illustrates this.

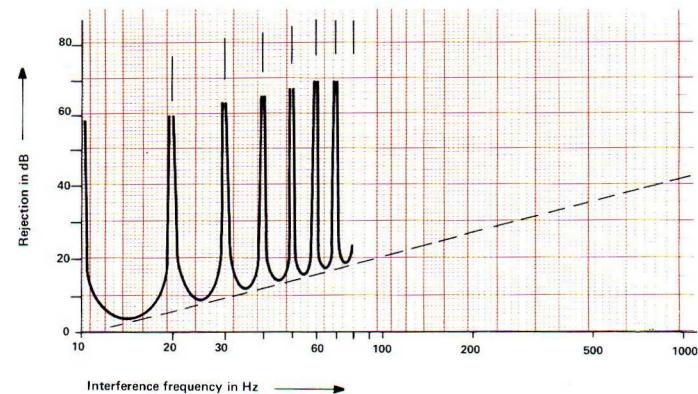


Fig. 5.2. SMRR as function of frequency in an integrating analog to digital converter.

Common-mode errors

Common-mode voltages are those voltages which appear on both sides of a signal line with respect to a common reference point, generally the common earth point. The common-mode rejection is generally specified separately for DC and AC voltages, and in most cases for an unbalance of 1 kΩ. The common-mode rejection for DC depends mainly on the insulation between the "low" and voltmeter ground. As can be seen in fig. 5.3, the voltage between high and low of the instrument's input is equal to the voltage drop across the 1 kΩ unbalance resistance, which forms a voltage divider together with the insulation resistance between "low" and ground. If the insulation resistance is increased, the voltage across the 1 kΩ resistor will be lower, and so will the input voltage of the voltmeter; so good common-mode rejection is highly dependent on the insulation resistance between "low" and ground.

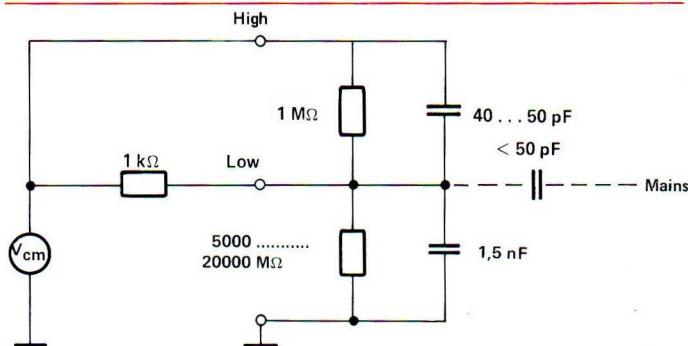


Fig. 5.3. Input circuitry of a DVM, showing the common-mode voltage V_{cm} .

The capacitance between "low" and ground is also of importance for AC voltages so it will be clear that the common-mode rejection is frequency-dependent. For good common-mode rejection for AC signals the capacitance between "low" and ground should be as low as possible. In general, this capacitance is given a value of the order of 1.5 nF – also for safety reasons. With this capacitance it is possible to obtain a common-mode rejection ratio (CMRR) for AC of about 50 dB at 50 Hz. If much higher CMRR's for AC are required, guarding techniques have to be used; see chapter 6.

The error caused by an AC common-mode signal in DC measurements is negligible. While the resulting AC signal at the input may be quite appreciable, the series-mode rejection of this signal leads to an over-all rejection of up to over 100 dB.

Sources of error in other than DC voltage ranges

The above considerations apply to DC voltage measurements. As far as digital multimeters are concerned, we also have to consider the accuracy of the other ranges. Because all parameters measured are first converted into DC voltages and each of these conversions involves a slight additional error, the accuracy for all other quantities will in general be less than for DC voltages.

In **DC current** measurements, it is the effect of the shunts that gives the additional error. First of all, the accuracy of the shunt resistance must be taken into consideration. Accuracies of better than 0.1% can be obtained, but are expensive. The temperature coefficient of this resistance also has a direct influence on the accuracy.

Secondly, the voltage drop across the shunt can also cause appreciable errors.

Let us suppose that a current of 150 mA flows from a 12 V supply source in the measuring circuit. The load resistance of the power supply is thus 80 Ω.

To measure the current, we insert the multimeter shunt into the measuring circuit. We assume this shunt to have a resistance of 1 Ω in the 200 mA range (this is a normal value). As a result, the current will change from 150 mA to $12 \text{ V} / 81 \Omega \approx 148.1 \text{ mA}$, which involves an error of more than 1.2%. For this reason, it may be more accurate to measure in the 2 A range (where a shunt resistance of 0.1 Ω would be normal), which only gives an additional error of about 0.12%. Theoretically, not taking other errors into account, this input current of 150 mA would give the following readings on a 4½ digit display:

- 148.14 mA in the 200 mA range.
- 0.1498 A in the 2 A range.

The measurement of **AC currents** involves a double conversion, from current to voltage and then from AC to DC, each conversion giving rise to an additional error.

In addition, it should be noted that shunts generally have a limited frequency range because of parallel parasitic capacitances and series parasitic inductances, and are certainly not suitable for the higher current ranges at frequencies above some kHz.

The additional error in **resistance** measurement depends mainly on the accuracy and stability of the measuring current through the resistance to be measured. Stabilized current sources can be constructed with reasonable accuracies in the mA and μA ranges, but difficulties increase rapidly in the nA range. If we take 2 V as the maximum permissible voltage across the resistance to be measured, this means that the current in the 200 kΩ resistance

range will be no more than $\frac{2 \text{ V}}{200 \text{ k}\Omega} = 10 \mu\text{A}$,

while in the 20 MΩ range it will be 100 nA, and 1 nA in the 2 GΩ range.

This is why the accuracy of resistance measurements falls off rapidly in the MΩ ranges; it will be clear that the 1% accuracy at 2 GΩ as specified for e.g. the Philips PM 2527 is an exceptional achievement.

AC-DC conversion errors

The accuracy of AC voltage measurements depends in practice only on the accuracy of the AC-DC converter. The type of converter used is most important in this connection. Most multimeters on the market use average voltage detectors as described in chapter 4.

As long as purely sinusoidal signals are measured, excellent result can be obtained. However, with distorted signals this system can give rise to significant errors. True RMS AC-DC converters are thus to be preferred for high accuracy. Full details on the advantages of true RMS conversion are given in chapter 7.

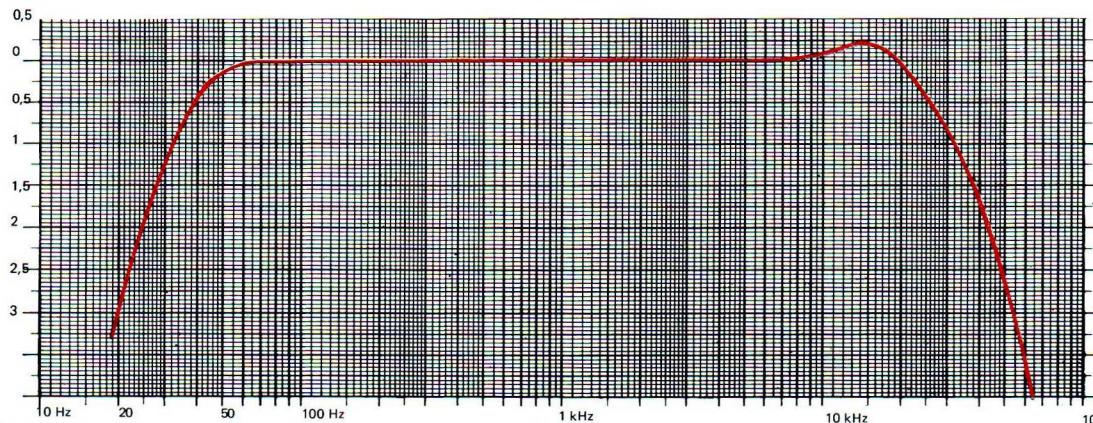


Fig. 5.4. Frequency characteristic of a DVM.

Frequency range

While the bandwidth of an oscilloscope is nearly always specified for a 3 dB drop in gain, no common standard exists for multimeter frequency ranges. A frequency range as illustrated in fig. 5.4. could be specified as e.g. Accuracy 0.2% of reading $\pm 0.1\%$ of range at reference frequency of 1 kHz.

Bandwidth:

46 Hz to 22 kHz for accuracy 0.2% of reading $\pm 0.1\%$ of range.

40 Hz to 25 kHz for accuracy 0.5% of reading $\pm 0.1\%$ of range

20 Hz to 55 kHz for accuracy 3% of reading $\pm 0.1\%$ of range

It is essential not to compare frequency figures only but to do this in combination with the accuracy specification.

Noise

Another source of error which is encountered mainly at the lower end of the ranges is noise. Generation of noise can be limited but never completely avoided. For example, the thermal noise of the input resistance gives rise to an error voltage in the AC ranges at open input. Fortunately, the influence of these noise voltages is slight. It can be estimated from the formula:

$$V = \sqrt{R^2 - S^2}$$

where V is the voltage applied to the input, R is the reading displayed and S is the error voltage of the instrument at the same source impedance as the circuit to be measured. With source impedance up to $10 \text{ k}\Omega$, most AC ranges will not give more than 1% of range end value error voltage.

At a reading of 10% of range, this means that the error due to noise is of the order of 0.5% of reading decreasing to 0.005% at full scale.

Conclusions

We have seen that the accuracies obtained in measuring practice depend not only on the specifications of the instrument but also on a number of secondary error sources in the measuring instrument or in the circuit under test. The secondary errors can make up a large part of the total error, especially when the intrinsic accuracy of the instrument is high. Fortunately, most of these errors are strongly dependent on the source impedance; as long as this impedance is less than $10 \text{ k}\Omega$, their influence is fairly limited. For source impedances above $10 \text{ k}\Omega$, however, we certainly have to take the additional errors into account. We may state as a general rule that if laboratory DC measurements are performed with an average error of about 0.1 to 0.2%, the technicians are doing an excellent job. For AC measurements, it is difficult to obtain an average error of less than 1%.

The pros and cons of digital voltmeters

We will conclude this chapter of the digital instrument course with a brief survey of the strong and weak points of digital voltmeters.

There are several reasons for choosing a digital meter for a given measuring application. The main ones are:

- The high accuracy that can be obtained (notwithstanding the limitations discussed above).
- The unambiguous reading obtained, which is very useful when large series of measurements have to be performed or technicians without a great deal of training have to carry out the measurements.

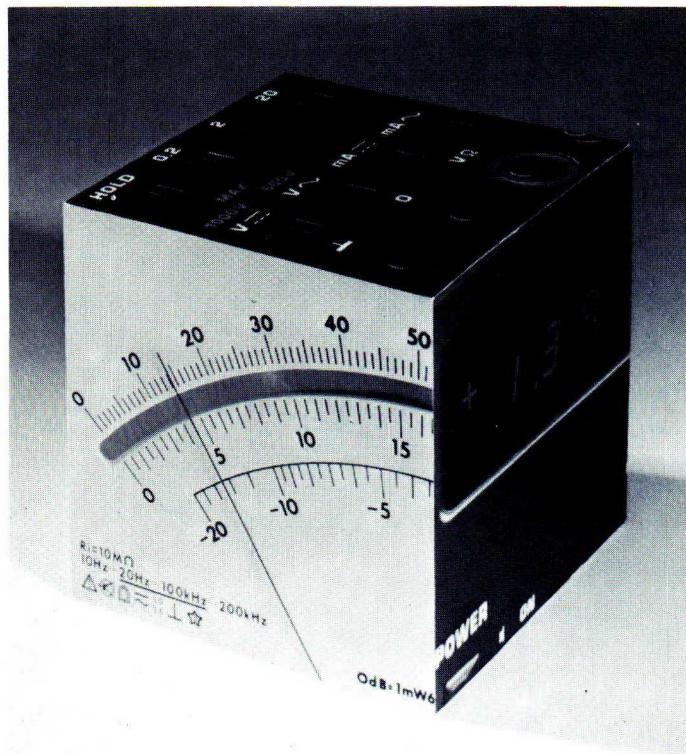
- The high resolution (1 part in 10^3 or 10^4 or even better), which is particularly useful for the measurement of small differences or stabilities.
- The possibility of full automation, making measurement a simple touch-and-look operation.
- Last but not least, the possibility of using these instruments (when provided with a digital output) in automatic test systems for data acquisition or even more complex data handling.

On the other hand, there are a number of application for which digital equipment is less suitable:

- When high-sensitivity measurements have to be made, e.g. under 1 mV for AC and under $100 \mu\text{V}$ for DC, noise and drift will cause the last digit of the display, and possibly even the last but one, to fluctuate a lot; it is difficult to estimate an average reading under these conditions.
- Adjustment of the reading to certain value, and in particular to a minimum or maximum, is by no means easy with a digital instrument.

In these two cases, a good analog instrument makes the operation much easier.

The introduction of micro-processors in digital voltmeters may eventually provide a solution for these two problems, apart from the many other undoubted advantages micro-processors have to offer.



Questions

Check the answer(s) you believe to be correct.

Q.5.1. A $4\frac{1}{2}$ -digit autoranging DVM has a specified accuracy of $\pm 0.05\%$ of reading $\pm 0.02\%$ of range. Its maximum range is 2000 V. When a voltage of 220 V is measured, the overall accuracy will be

A 0.07%
B 0.23%
C 0.51%

| | |
|---|--|
| A | |
| B | |
| C | |

Q.5.2. What is the load error when the voltage across a $5\text{k}\Omega$ resistor is measured with a $4\frac{1}{2}$ -digit DVM with an input impedance of $10\text{M}\Omega$?

A 0.01%
B 0.02%
C 0.05%

| | |
|---|--|
| A | |
| B | |
| C | |

Q.5.3. A DVM measures the current through a 1000Ω resistor, with the range selector of figure 4.3. set to 200 mA. The error introduced by connecting the meter in series with the resistor is

A 1 %
B 0.1 %
C 0.01%

| | |
|---|--|
| A | |
| B | |
| C | |

Q.5.4. Which of the following statements are true ?
A DVM has the following advantage(s) over an analog voltmeter:

A a higher resolution can be obtained
B voltages below $100\mu\text{V}$ can be measured with greater accuracy
C it is more convenient for monitoring the adjustment of a voltage to a maximum or minimum value

| | |
|---|---------------|
| A | true/not true |
| B | true/not true |
| C | true/not true |

The answers to the problems will be found at the end of the book.

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PROBE

GUARD

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V

Chapter 6

Guarding techniques

Introduction

Measuring systems involving instruments such as digital voltmeters and recorders with accuracies of around 0.1% or even better must be designed with great care to keep parasitic voltages and other electrical noise out of the system. For example, the recorder PM 8222 has a guaranteed accuracy of 0.25%. When this instrument is used with a full-scale deflection of 1 mV, a parasitic voltage of $10 \mu\text{V}$ (which can easily arise in the measuring system unless special precautions are taken) represents an error of 1%, thus completely annulling the high accuracy which is one of the main features of this instrument.

One of the main sources of parasitic voltages in measuring systems is the mains voltage, and in this chapter we shall discuss mainly means of reducing errors due directly or indirectly to the mains voltage. There are two main methods of eliminating these errors: the use of isolating transformers and the shielding of leads. The shielding of leads helps to keep out induced parasitic voltages. This form of "electrical environmental pollution" has assumed considerable proportions now that so much electrical power equipment discharges its operating current to earth.

However, the use of shielding techniques and isolating transformers can itself lead to errors in the measuring system. The main purpose of this chapter is to explain how this can come about, and how guarding systems can be intelligently designed on this basis to keep out parasitic voltages.

Distinction between safety guard and signal ground

Many instruments have three input terminals, marked "High", "Low" and "Earth" or "+", "-" and "=". The "-" or "+" terminal is the signal ground. Both the biasing voltages and the voltages to be measured in the instrument are related to this. The terminal marked "=" is the safety earth; all external metal parts of the instrument are connected to this point, which is in its turn connected to the earth lead of the AC mains cable.

In a proper installation this earth connection ensures that none of the conducting parts of the instrument which personnel or other people could touch can rise in potential above the safety margin (42 V) with respect to earth. A metal link is often provided between the "-" (or "Low") and "=" terminals (fig. 6.1.). The signal "common" point is then earthed. If the instrument in question has the only connection to earth in the measuring system, measurements can be performed in a straightforward way. However, measuring errors may occur in cases where the device under test has also a connection to earth. (See the section "Ground loops and ground currents" next page). The terminals "Lo" and "=" must now be disconnected. The instrument's cabinet is of course still connected to earth, so that the system is still electrically safe, but the above-mentioned errors are eliminated because the "common" is floating with respect to earth. The reason for this will be considered in the next section.

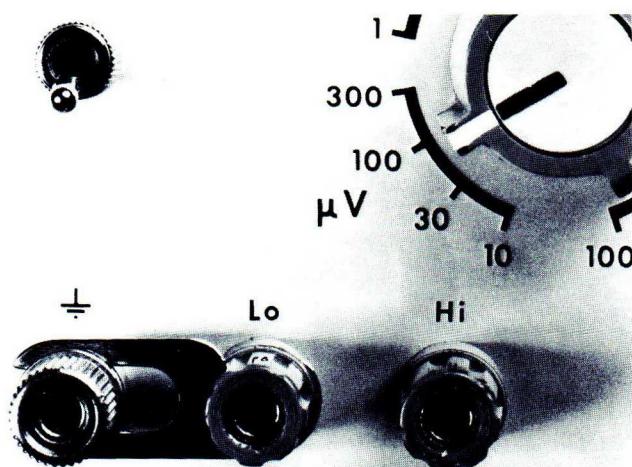


Fig. 6.1. A conducting link between the "Lo" and "=" terminals used to ensure safe earthing of the housing in a PM 2434 DC microvoltmeter

Ground loops and ground currents

We all know that the metallic cabinet of electrical equipment should be earthed as mentioned above, to ensure electrical safety. In power installations, the earth connection may be obtained by driving a copper or iron rod into the ground to a depth of 10–100 feet beneath the surface. The depth will depend on the local conductivity of the soil, which is subject to seasonal variation.

If one rod does not give a sufficiently low earth spread resistance, a system of rods connected by a copper ground bus can be used. The required value of the earth spread resistance depends on the maximum current which may be obtained from the power system in use. In any case, possible defects in the equipment should not lead to a voltage exceeding 42 V between the metallic housing of an instrument and earth.

Sometimes the iron rods of a reinforced concrete structure are welded together and connected to special conducting rods to give the required low earth spread resistance. A metallic underground water piping system also provides good contact with earth and is probably the best solution for the earthing of measuring instruments and similar equipment, if available.

As we mentioned in the introduction, considerable ground currents may be expected as a result of the discharge of current from power systems to ground. It is very difficult to evaluate these currents and no satisfactory methods of measurements have been developed to date. Most of the data available are average values taken over a period of time because of the complexity of parameters such as soil resistivity, soil moisture, depth of the electrodes, weather conditions etc.

Now, when two points in e.g. a measuring system are connected to earth, we get what is known as a "ground loop" formed by the part of the measuring system between the two points and the soil between the two earth points. Fig. 6.2 shows an example of such a ground loop formed between an earthed transducer and an earthed measuring instrument which are connected by relatively long cables. R_1 and R_2 represent the cable resistances. The transducer might be a thermocouple or a strain gauge connected to an earthed object, while the instrument can be a strip chart recorder, an impedance bridge, an oscilloscope or a data acquisition system.

The earth currents and the soil resistivity create a virtual voltage source E_{cm} , causing a current to flow through the measuring set-up. The measures to be taken in order to avoid or to reduce the effect of this current will be discussed below.

An analogous situation arises in a measuring installation consisting of different instruments mounted in a 19" rack.

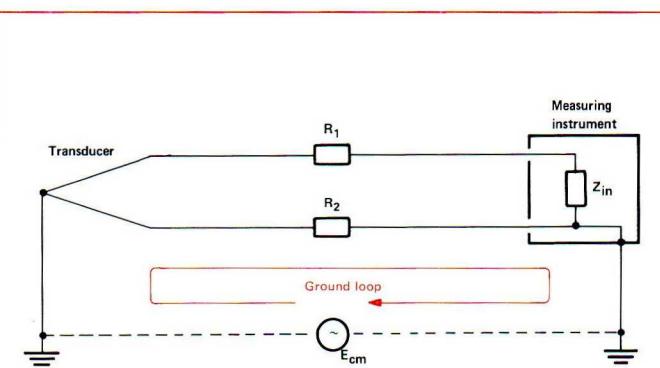


Fig. 6.2. Formation of a ground loop

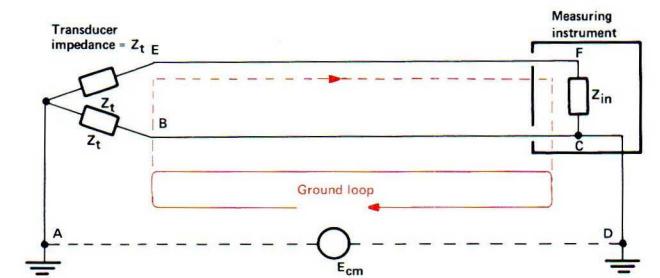


Fig. 6.3. Conversion of a common-mode voltage into a series-mode voltage

A current flow through the rack via the signal ground terminals of the instruments, no matter whether the latter are connected to earth or not. If such a current does exist it is very likely to be coupled back into the circuit by means of resistive or capacitive coupling or electromagnetic induction. The current loop formed in this way is commonly called a "ground loop" too. The precautions which may be taken to reduce errors due to this type of ground loop will also be discussed below.

Summarizing, we may state that accurate measuring systems must be designed to reduce or eliminate errors due to:

- 1. ground currents flowing below the surface of the earth and
- 2. ground currents associated with the measuring set-up itself.

Common-mode and series-mode voltages

Common-mode voltages are those voltages which appear on both sides of a signal line to a common reference point, normally the common point or earth. As already mentioned above, they can be coupled to the measuring set-up in a resistive, capacitive or electromagnetic way. Such common-mode voltages can give rise to "series-mode" (normal mode) voltages in a measuring system. There are the series-mode voltages that really concern us, because they can lead to measuring errors.

The conversion of a common-mode voltage into a series-mode voltage is illustrated in fig. 6.3. In this measuring set-up a ground current flows in the low-ohmic circuit ABCD due to the common-mode voltage E_{cm} . However, a small current also flows through the transducer impedances Z_t and Z_{in} of the measuring instrument (AEFCD). Because Z_{in} is generally very high ($> 10^6 \Omega$), the current through AEFCD will be much smaller than that through ABCD ($Z_t \leq 1 \text{ k}\Omega$). A voltage will thus exist between the points E and B due to the difference in loop currents originating from E_{cm} . This voltage can be considered as being in series with the voltage which we intended to develop in the transducer, and has thus received the name "series-mode voltage" (also called "normal-mode voltage").

It is this series-mode voltage which causes us most concern, since it directly influences the accuracy of the measuring system. We will see that there are means available (filtering, integrating) for solving this problem to a great extent.

Both in fig. 6.2 and fig. 6.3 the source of E_{cm} is a difference in potential between the two earth points; this kind of common-mode voltage is therefore referred to as an *earth common-mode voltage*. It is generally related in frequency to the mains power supply.

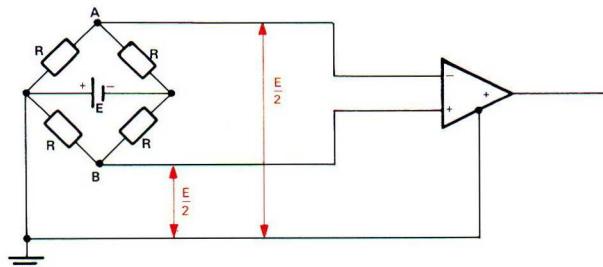


Fig. 6.4. Another kind of common-mode voltage source

Fig. 6.4 illustrates another kind of common-mode voltage source. Here the resistance bridge is fed by a voltage source E one terminal of which is earthed. The bridge could be of a normal resistance type but could also be a configuration of transducers. With one terminal of the supply voltage earthed, the bridge points A and B are at a potential of $E/2$ above ground. Thus the common-mode voltage at the input terminals of the amplifier is $E/2$. The power supply for the bridge is generally DC, but AC bridge supplies do exist too. This type of common-mode voltage is referred to as a *transducer common-mode voltage*. At the present state of the art, amplifiers and similar circuits will reject DC common-mode voltages up to 100–120 dB, but the possibility of error from this source should be borne in mind when extremely accurate measurements are to be made.

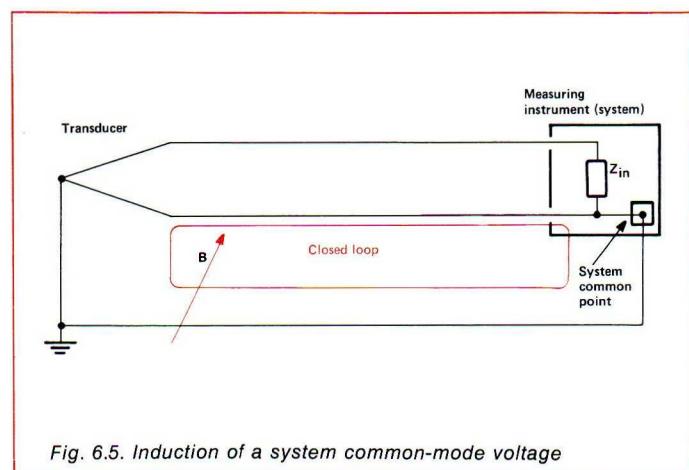


Fig. 6.5. Induction of a system common-mode voltage

The last type of common-mode voltage to be mentioned here is the *system common-mode voltage*, which is magnetically induced in a closed loop in the measuring system by an alternating (stray) magnetic field. Fig. 6.5 gives an example of a measuring set-up in which a system common-mode voltage may arise. The set-up is basically the same as in fig. 6.2, but in order to avoid earth common-mode voltages in the measuring system, the signal common points are combined in the system common point which is earthed at the transducer earth point. It will be clear that a voltage may be induced in the closed loop shown in the figure by a magnetic field B . The induced voltage is proportional to the flux density, the rate of change of flux and the circuit area. Therefore, the larger the area the more susceptible the system will be to magnetically induced common-mode voltage.

Keeping parasitic voltages out of measuring leads

Once we know the origin and kind of parasitic voltages likely to influence a measuring system, it will not be too difficult to understand the steps to be taken to reduce their effect. The two main methods used are *shielding* the signal path against intrusion of a parasitic voltage, and *isolating* this voltage from the signal path.

We shall first discuss some consequences of shielding a cable, and then the use of transformers to avoid hum in the measuring circuitry.

Shielding techniques

Shielding may be necessary to keep parasitic signals out of certain areas in low-level instrumentation. Before we can shield effectively against these signals, we have to know whether they are induced by electrostatic or electromagnetic fields. Electrostatic fields form a more common threat to electronic instrumentation, but are simpler to shield against: a metal housing will shield an instrument effectively against electrostatically induced currents because they flow through the surface. A braided copper sheet is most commonly applied for shielding cables, although aluminium foil is becoming increasingly important.

High-permeability ferromagnetic materials are the best shield against magnetic flux. These shields are for instance used round CRT's in oscilloscopes. The higher the permeability of a magnetic shield the better. Nickel-iron alloys have relative permeabilities in the range from 10 000 to 100 000.

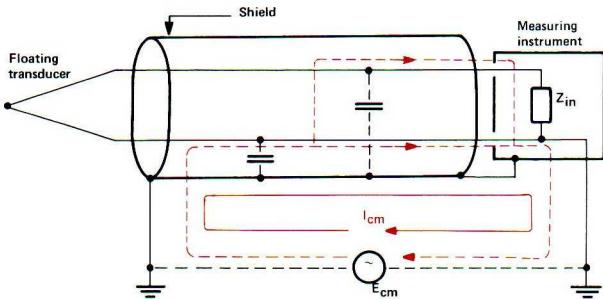


Fig. 6.6. Introduction of common-mode current I_{cm} in signal path due to incorrect grounding of shield

Regardless of the type of shield used, it is possible to cause more noise in the signal lines by improper grounding of the shields than would be present without shielding. Some examples will be given here. In fig. 6.6 a common-mode current I_{cm} is introduced in the signal path by capacitive coupling to the shield which is connected to earth both at the transducer side and at the instrument housing. The transducer is floating here.

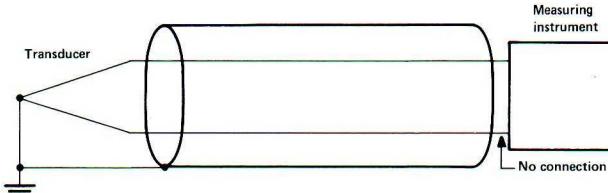


Fig. 6.7. A better way of grounding the measuring set-up of Fig. 6.6

Figure 6.7 gives a better solution in this case, with an earthed transducer and a floating instrument. When the transducer consists of an ungrounded bridge configuration the whole bridge may be shielded and earthed. In many signal transport configurations, line amplifiers have to be used to keep the signal at a proper level. In case of a single-ended amplifier, the system should be shielded as shown in fig. 6.8.

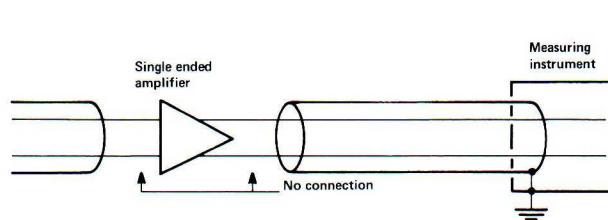


Fig. 6.8. The right way to ground a measuring set-up with a single-ended amplifier

It will be seen that the amplifier is not connected to the shield. However, there are right and wrong ways of applying this shielding technique. Fig. 6.9 shows one of the wrong ways. Obviously, the two ground-loop currents can introduce series-mode voltages in the measuring chain by capacitive coupling.

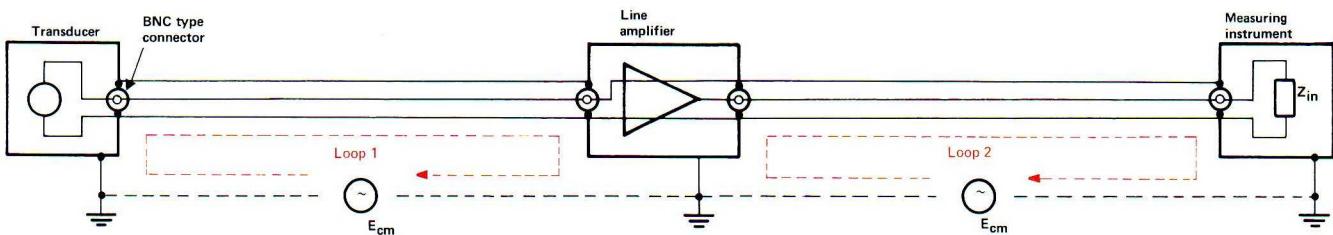


Fig. 6.9. An incorrect application of the principle of Fig. 6.8

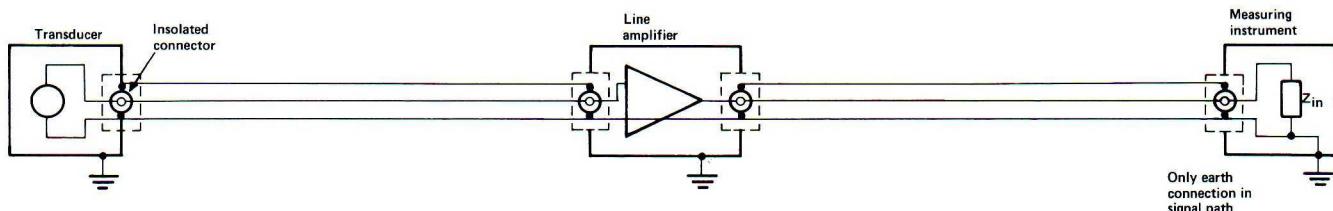


Fig. 6.10. A correct application of the principle of Fig. 6.8

A correct way of connecting the shield for this set-up is given in fig. 6.10. Although all instrument housings are earthed for safety reasons, no ground current can flow because the only point where the shielding has been earthed is at the instrument side.

Power isolation

The first step in reducing parasitic voltages derived directly from the mains supply is to apply a power isolation transformer. Such a transformer is built into most instruments. The reason for this is illustrated in fig. 6.11. The instrument is connected to the mains via a 3-core cable. Of the two leads carrying the power current, one is called the neutral lead and is connected to earth at the primary power distribution point such as a step-down transformer.

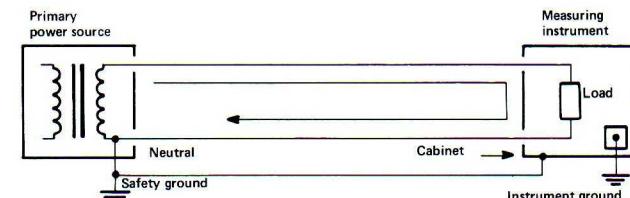


Fig. 6.11. The reason why a power-isolation transformer is included in most measuring instruments (see text)

However, the safety guard wire is also earthed at the same point to which all metal housings are connected via the third lead of the mains cable. This provides a low-ohmic path to earth for currents caused by insulation breakdown, accidental shorts etc. A common-mode voltage can be formed between this power ground and the instrument ground when the two ground systems are connected, and this may give a large amount of hum in the equipment.

A power isolating transformer can eliminate most of this common-mode voltage. The shields around the primary and secondary windings are connected to one side of the winding, so that stray capacitances between the windings and the shields are eliminated at one end. Moreover, a large capacitance is achieved between the shield and the cable core, giving a low-impedance path for the common-mode generator E_{cm} . The secondary shield is often connected to a centre tap of the secondary winding, fig. 6.12.

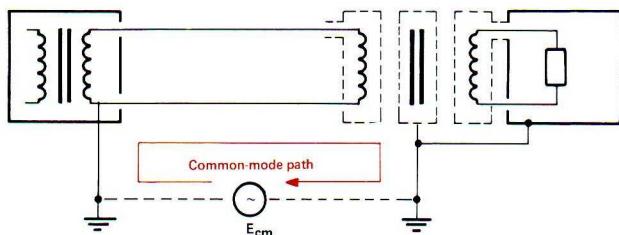


Fig. 6.12. Principle of shielding a measuring instrument with a power-isolation transformer

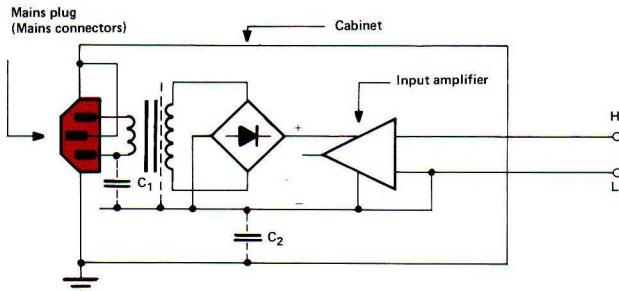


Fig. 6.13. Application of the shielding principle of Fig. 6.12 in the digital multimeter PM 2421

Note that this also reduces the stray capacitances between the primary and secondary windings to a value of the order of 0.1 pF, while *without* shielding the capacitive coupling between the two windings may be of the order of 2000–5000 pF. Figure 6.13 shows how this principle is applied in the PM 2421 digital multimeter. The specifications for this instrument are:

Cap. mains-L_o : 50 pF (C₁ in fig. 6.13)

Cap. L_o-chassis: 1,5 nF (C₂ in fig. 6.13).

It will be clear that 50 pF gives quite good separation between the mains and the measuring circuitry.

When even more stringent input specifications (in particular better separation between measuring circuitry and mains) are needed, a solution like that shown in figure 6.14 is required. This is used in Philips' multipoint recorder PM 8235. An extra transformer T₂ has been added here to reduce the influence of the mains. One of the secondary windings of T₁ is used to feed a 100 kHz oscillator which forms part of the measuring system. The output of amplifier A is fed to the remaining (less sensitive) part of the instrument's amplifier via a galvanically separated, light-coupled diode device. A floating guard, which may be connected to an extra terminal on the instrument, is placed around the input amplifier A and its supply unit to eliminate the influence of external parasitic voltages. This set-up gives an AC common-mode rejection ratio (CMRR) of 126 dB for the mains frequency.

Common-Mode Rejection Ratio (CMRR)

So far we have been considering how to reject parasitic voltages entering the measuring instrument. However, the shielding at the instrument side can also be designed so as to improve the CMRR. This will be explained with reference to figures 6.15 and 6.16, in which practical values of the impedances are assumed. The instrument is a DVM, the input amplifier of which is floating, i.e. it has no galvanic connection to earth. However, although there is no direct connection between the input amplifier and ground, an AC common-mode current I_{cm} will cause a series-mode voltage across R_t via a ground loop through C₁.

Example: If C₁ = 10 000 pF and R_t = 1000 Ω then the following calculation can be made

$$X_{C_1} = \frac{1}{2 \pi f C_1} \approx 300 \text{ k}\Omega \text{ at } 50 \text{ Hz}$$

If we neglect the current through the very high input impedance of the input amplifier (see also Fig. 6.3.), then the part of E_{cm} which appears across R_t as series-mode voltage can be calculated to be

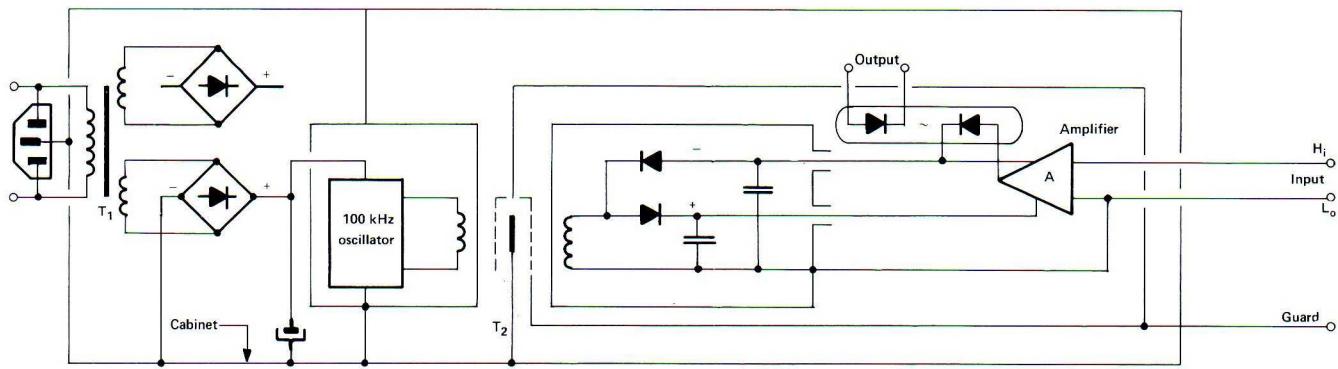


Fig. 6.14. Application of the shielding principle of Fig. 6.12 when more stringent input requirements have to be met as in the multipoint recorder PM 8235

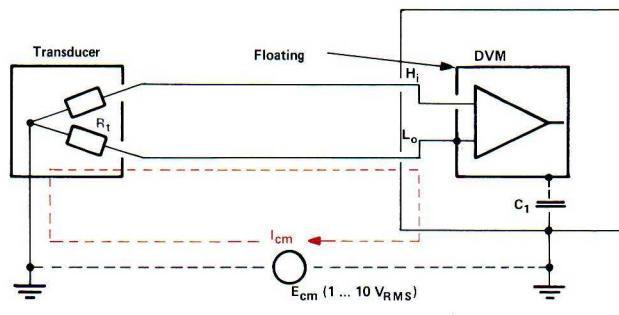


Fig. 6.15. The CMRR of this measuring set-up can be improved by reducing the value of C_1

$$V_{R_t} = \frac{R_t}{R_t + X_{C_1}} E_{cm}$$

For E_{cm} 3 V we find

$$V_{R_t} \approx \frac{1}{300} 3 \text{ V} = 10 \text{ mV}$$

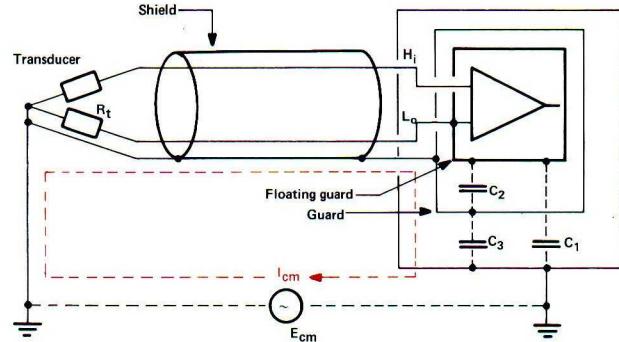


Fig. 6.16. The CMRR of the set-up of Fig. 6.15 can be improved still further by inclusion of a second guard

So in this particular case C_1 and R_t determine the amplitude of the series-mode voltage. The floating input thus reduces the interference only to the extent that

$$X_{C_1} = \frac{1}{2 \pi f C_1} \text{ is large compared with } R_t.$$

In DC voltmeters, integration techniques and filtering can reduce the series-mode voltage. If the DVM in question has been designed to measure AC voltages, the above 10 mV will be added directly to the display as an error. The **Common Mode Rejection Ratio (CMRR)** is therefore defined as the ratio of the measuring error due to the common-mode voltage and the common-mode voltage itself:

$$\text{CMRR} = \frac{E_{\text{cm}}}{E_{\text{cm}} \text{ measured}}$$

or in dB

$$\text{CMRR} = 20 \log \frac{E_{\text{cm}}}{E_{\text{cm}} \text{ measured}} \text{ (dB)}$$

In the above example E_{cm} measured is 10 mV and $E_{\text{cm}} = 3 \text{ V}$, so

$$\text{CMRR} = 20 \log \frac{3000}{10} \approx 50 \text{ dB}$$

From this example we see the important fact that the CMRR depends on:

- the frequency of E_{cm}
- the value of C_1 ,
- the transducer impedance R_t

Thus is not sufficient only to specify a CMRR of 50 dB. The conditions under which the CMRR has been defined must also be stated. These conditions are generally standardized e.g. by the International Electrotechnical Committee (IEC, subcommittee 13B, draft April 1972, section 23.5).

Floating guard

It will be clear from the previous section that one way of improving the CMRR is to increase the value of X_{c_1} or, in other words, to diminish C_1 in figure 6.15. A second guard in the DVM improves the above result very appreciably, see figure 6.16. (In this way we obtain much the same configuration as in figure 6.14.) The measuring cable of figure 6.16 has been provided with a shield, which is connected to the guard at the instrument side and to earth at the transducer side.

The common-mode voltage E_{cm} will again produce a current I_{cm} . At the connection of the transducer tip and shield this current I_{cm} will split up in two parts:

- one part will flow via the shield and C_3 back to E_{cm} without effecting the measurement and
- the other part will flow through the lower R_t , the L_o terminal C_2 and C_3 back to E_{cm} .

Because $I_{\text{cm}} R_t$ is very small compared with E_{cm} (owing to the impedances in this loop), only a very small voltage is developed across both plates of C_2 which are thus at approximately the same potential. Hence the current through C_2 will be very small. This means that the effective capacitance of C_2 will be of the order of magnitude of some pF, although C_2 itself can be as high as some nF. Compared with the value of 10 nF in the above example this means an improvement of another 60 dB for $C_{2\text{eff}} = 10 \text{ pF}$.

One could consider this effect in a slightly different way: the guard is connected via the shield to the same potential as the L_o terminal via R_t so that no current will flow through the capacitance between L_o and GUARD (C_2). It should be borne in mind that both the above considerations are only valid as long as $R_t \ll X_{c_2}$. Thus we must keep the impedance between L_o and GUARD high; it would be wrong to reduce this impedance on the grounds that there is no current through it anyhow.

In figure 6.16. the effective capacitance between L_o and EARTH is symbolized by C_1 and is in the range of 0.5–5 pF. The capacitance between GUARD and EARTH is of the order of some nF.

It is obvious that the capacitances C_1 , C_2 and C_3 in figure 6.16. are bypassed by the insulation resistances. A complete equivalent circuit is given in figure 6.17.

The practical values of the impedances have been taken from the DC DVM PM 2441 with only one exception:

$$C_1 = C_{1\text{eff}} = 3 \text{ pF}$$

With the values given in the figure, we can calculate the series-mode voltage developed across R_t .

$$\text{For } 50 \text{ Hz: } X_{c_2} = \frac{1}{2\pi 50 \times 700 \times 10^{-12}} \approx 4.5 \text{ M}\Omega$$

thus $X_{c_2} \ll R_t$ and $X_{c_3} \ll R_3$ at mains frequency.

For X_{c_1} we find $X_{c_1} \approx 1 \text{ G}\Omega$ which is still much smaller than $60 \text{ G}\Omega$ the insulation resistance between L_o and EARTH. So we can calculate I_{cm} from the capacitances only.

Now, again neglecting the current through R_{in} of the input amplifier and bearing in mind that no current flows through C_2 , we find a series-mode voltage V_{R_t} developed across R_t of

$$V_{R_t} = \frac{R_t}{R_t + X_{c_1}} E_{\text{cm}}$$

With $R_t = 1 \text{ k}\Omega$ and $X_{c_1} = 1 \text{ G}\Omega$ we find a CMRR for 50 Hz of

$$\text{CMRR} = 20 \log \frac{R_t + X_{c_1}}{R_t} \approx 120 \text{ dB}$$

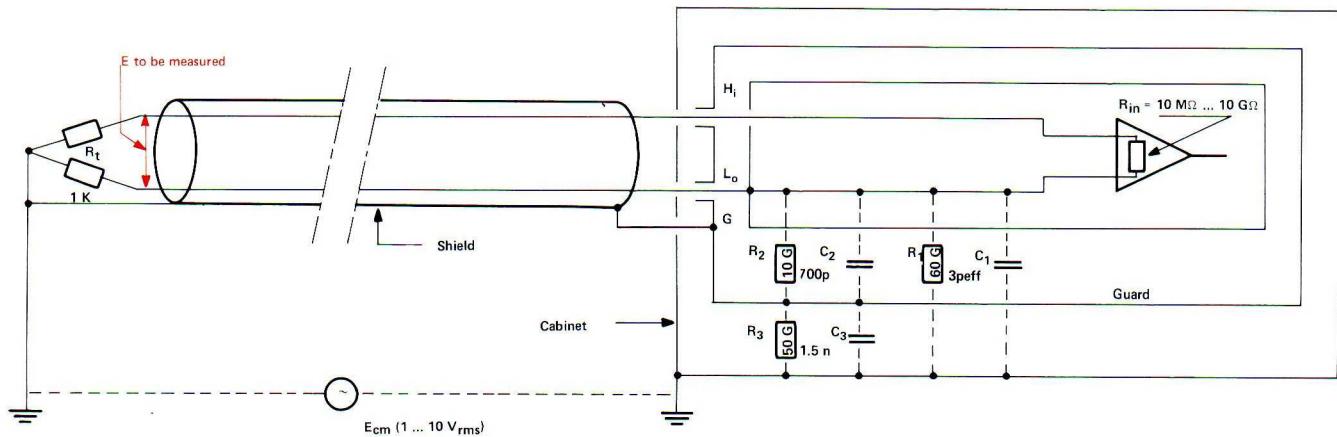


Fig. 6.17 Equivalent circuit of the set-up of Fig. 6.16.

Note: If we did **not** have the guard connection, then we would have to reckon with a current through C_2 and C_3 in series (≈ 480 pF) and the CMRR would be 76 dB.

For DC the common-mode current which influences the measurements flows through R_t and R_1 . With the values of figure 6.17 it can be calculated that for DC the CMRR will be as high as 155 dB. This is due to the high insulation resistance of 60 G Ω compared with an R_t of 1 k Ω .

Why are these figures of such importance? As already stated in connection with fig. 6.15, the instrument in use may be a DVM. When this instrument has an accuracy of 0.001% or less and a resolution of $1\mu\text{V}$ then a DC series-mode voltage of $50\mu\text{V}$ is extremely important, because the last two digits are meaningless when the DVM is used in its lowest range of 1 V. An effectively guarded system, though expensive, is thus essential for high accuracy and resolution in low-level measuring instruments.

An AC series-mode voltage also influences an extremely accurate reading of the DVM. But as the AC common-mode voltage is generally at mains frequency, it depends on the nature of the voltmeter's analog-to-digital converter to what extent this small error can be further reduced (see chapter 2).

Doubly-shielded cables

The floating shield may also be used the other way round: in order not to load a signal source when an AC voltage has to be measured accurately. Then the principle indicated in figure 6.18 is used. Via the input amplifier the floating inner shield of a doubly-shielded probe cable of an AC voltmeter is brought to exactly the same potential as the voltage E to be measured. C_1 is thus made ineffective because no current will flow through it. The capacitive load of the cable is now formed by C_2 , but the current through it is delivered by the input amplifier via its output A and not flowing through R , thus causing no series-mode error. Obviously the input resistance of the amplifier is very high.

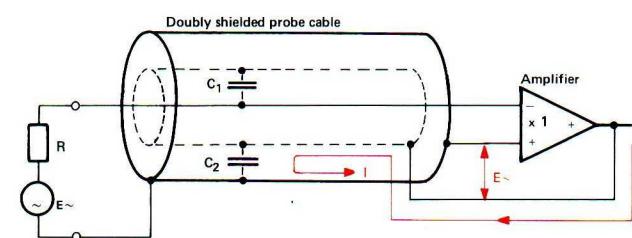


Fig. 6.18. Principle of double shielding of a probe cable

An application

To demonstrate the importance of a good guarding technique, an example is given below.

Figure 6.19. shows an automatic measuring set-up consisting of a scanner PM 2460/2461, a DVM (PM 2442) and a printer (PM 2466), which can be used to measure voltages at n different points (up to 1000). The voltage at each point is connected to the respective input of the scanner, channel 1... n . Depending on the setting of the control unit the channels are scanned one by one after a start command. First the GUARD is connected to the DVM by a 3-pole relay. 1 ms later the control unit energizes the 2-pole relay so that now H_i and L_o are connected to the DVM. An adjustable time (between 5 and 50 ms) after the latter relay contacts have closed, the control unit supplies a start pulse to the DVM. As soon as a channel is measured, the DVM generates a command for the printer to print out the measured value.

During printing, the printer supplies a HOLD OFF signal to the scanner, causing the relays to be de-energized while the scanner is clocked to switch on the next channel. As soon as printing is finished, the HOLD OFF signal changes state and the relay contacts of the next channel are closed, depending on the setting of the control unit. Note the extra precautions taken to ensure a good GUARD connection. It would be much easier and cheaper to use only the 2-pole relay, but then the guards (shieldings) of all the measuring points would be connected together at the scanner side. It is clear from the above that this procedure would simply be asking for measuring errors. The $3.03\text{ k}\Omega$ resistor in the guard line of the relay unit PM 2461 has been used for current limitation in case of different common-mode voltages at the measuring points.

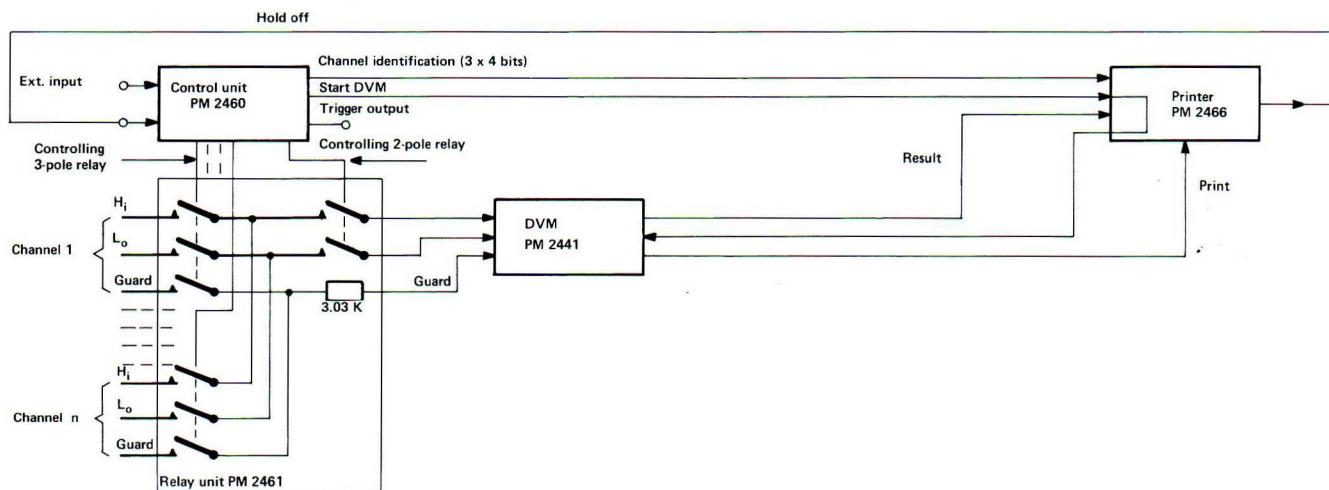


Fig. 6.19. An example of proper shielding of an automatic measuring set-up

Questions

Check the answer(s) you believe to be correct.

Q.6.1. The common-mode voltage E_{cm} of fig. 6.20 is converted into a series-mode voltage because

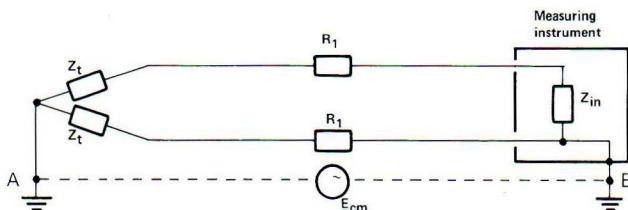


Fig. 6.20

- A points A and B do not have the same potential
- B the transducer impedance $Z_t \gg$ the cable resistance R_1
- C the measuring instrument's input impedance $Z_{in} \gg (Z_t + R_1)$

| | |
|---|--|
| A | |
| B | |
| C | |

Q.6.2. The common-mode voltage at both input terminals of the operational amplifier in fig. 6.21 is

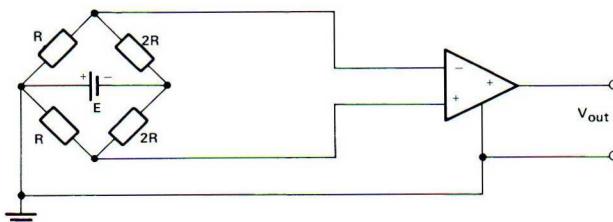


Fig. 6.21

- A E volt
- B $E/2$ volt
- C $E/3$ volt

| | |
|---|--|
| A | |
| B | |
| C | |

Q.6.3. The bridge voltage in fig. 6.21 above is $E = 3$ V. The CMRR of the op-amp for DC is 100 dB. With the Wheatstone bridge in equilibrium, the output voltage V_{out} is

- A 0 V
- B 10μ V
- C 15μ V

| | |
|---|--|
| A | |
| B | |
| C | |

Q.6.4. The influence of E_{cm} in the measuring circuit of fig. 6.22 can be eliminated by:

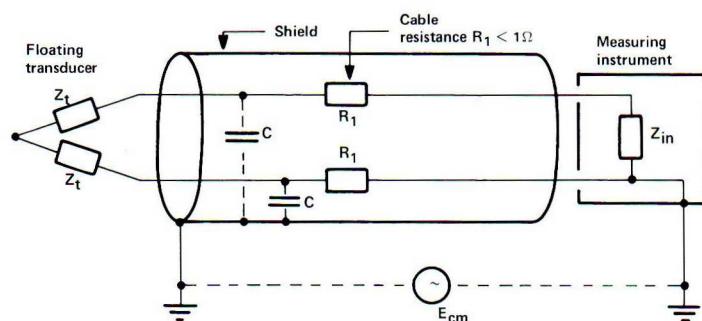


Fig. 6.22

- A disconnecting the shield from earth at the transducer side
- B disconnecting the shield from the instrument
- C making the instrument's input impedance $Z_{in} = \infty$

| | |
|---|--|
| A | |
| B | |
| C | |

Q.6.5. In the DMM of figure 6.23 the stray capacitance between the L_o terminal and the cabinet is specified as $\leq 2.6 \text{ nF}$. The transducer impedance is $1 \text{ k}\Omega$. The CMRR for AC measurements in a country with 60 Hz mains supply will be

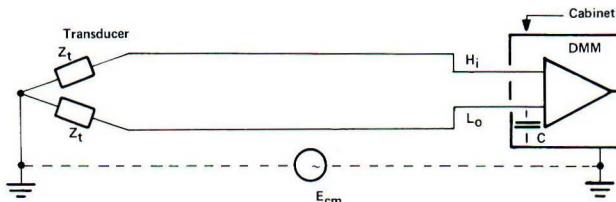
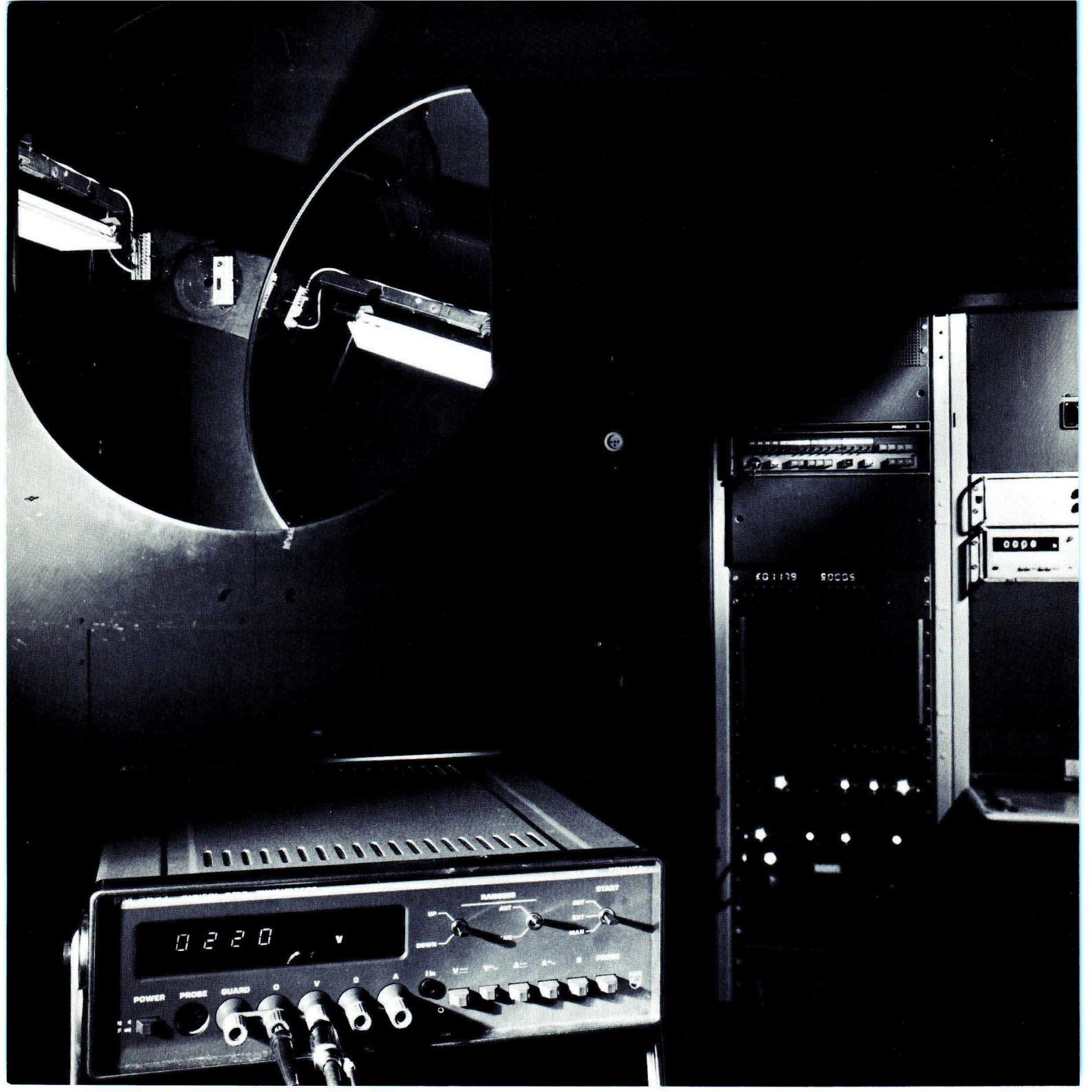


Fig. 6.23

- A 50 dB
- B 60 dB
- C 80 dB

| | |
|---|--|
| A | |
| B | |
| C | |

The answers to the problems will be found at the end of the book.



0220

POWER NOISE GUARD

RANGE
SWITCH
START

A B C D E F G H I J K L M N O P Q R S T U V W X Y Z

401129 50005

Chapter 7

AC and RMS measurements

Introduction

Practically all AC voltage and current measurements today are performed with electronic instruments which in fact measure either the average or the peak value of a sine wave, while displaying the RMS value. There is no harm in this as long as we are dealing with pure, undistorted sinusoidal waveforms. Unfortunately, however, sine waves are generally distorted in practice, which causes the above-mentioned instruments to have a faulty readout. This is one of the reasons why the RMS meter was developed. This chapter gives the why and wherefore of conventional AC measurements and the definition of RMS; an interpretation of instrument specifications as regards RMS measurements and some applications of such measurements will be dealt with.

Hidden errors in conventional AC measurements

The method used for measuring alternating voltages in conventional electronic instruments is based on the assumption that only sine waves have to be dealt with. Because of the well defined shape of the sine wave, we can derive its RMS value from either the peak value or the average value, which are easier to measure. We see, from the text in figure 7.1., that if we detect the peak value of a sinusoidal voltage with a peak detector, we simply have to multiply by 0.707 to know the RMS value. The basic circuit of a peak detector is shown in fig. 7.2. Since the diode D is in the feedback loop of the operational amplifier, the forward voltage drop is divided by the open-loop gain. Once the diode is conducting the output voltage is multiplied by $R_1/R_2 = 0.707$ so the detected DC voltage E_{out} across C is equal to the RMS value of the input E_{in} .

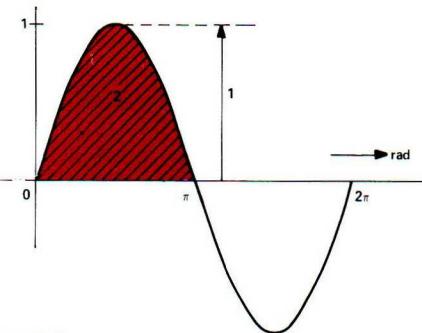


Fig. 7.1. Sine wave.

Mathematical calculations (see Appendix) show that if the amplitude of a sine wave is 1, then its RMS value is $1/\sqrt{2} \approx 0.707$.

If amplitude = 1, then the average value of a half period = $2/\pi$.

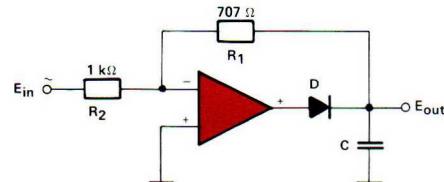


Fig. 7.2. Basic circuit of peak detector

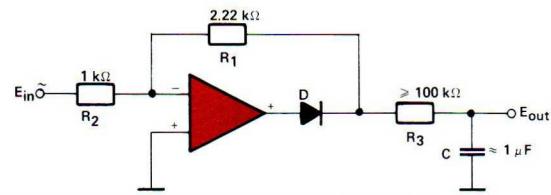


Fig. 7.3. Basic circuit for averaging 1/2 sine wave

A basic **averaging** circuit is shown in figure 7.3. It contains the same rectifying circuit as in figure 7.2, but now C is fed via a large resistor R_3 ; these two together form an integrating circuit. As we see from figure 7.1, the average value of half a sine wave is $2/\pi$ times the amplitude, so if we integrate half a sine wave over a whole period, the average will be half of this value, or $\frac{1}{2} \times 2/\pi = 1/\pi$ times the amplitude. Since the RMS value is $1/\sqrt{2}$ times the amplitude in this case we have to multiply by a factor

$$\frac{1/\sqrt{2}}{1/\pi} = \frac{\pi}{\sqrt{2}} \approx 2.22$$

As in figure 7.2, in figure 7.3 this is performed by R_1 and R_2 . Note that for a voltmeter which is designed to respond to a full-wave rectified sine wave, this factor becomes 1.11.

Which systematic errors can be hidden in these measuring systems?

It will be clear from figure 7.4. that in the presence of noise an AC voltmeter responding to the peak value of the input signal will give a measuring result equal to the amplitude of the sine wave plus that of the noise. If for instance the noise is $\pm 1\%$, the error of the measuring result will also be $\pm 1\%$; so measuring this with a **digital** instrument of 0.1% accuracy makes the last 2 digits meaningless.

On an oscilloscope, 1% noise is hardly visible, since 1% of say a 5 cm peak-to-peak display is only 0.5 mm which appears merely as a slightly thicker trace on the screen. An instrument responding to the average value of the input signal will display the correct value of sine wave in the presence of noise, since the noise will be averaged out. Of course, if we **want** to measure the noise accurately the RMS meter is the only possible choice: since the noise represents a certain power, an RMS meter will display the value of the signal plus noise applied to the input terminals.

Another type of "noise" frequently met with is spikes superimposed on the mains voltage. It will be clear from figure 7.5 that a peak-responding instrument will give enormous errors here. These spikes can occur when someone in the neighbourhood is controlling mains power by means of thyristors.

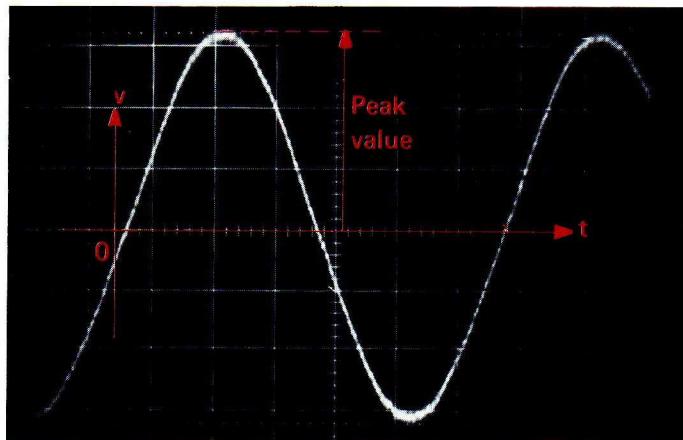


Fig. 7.4. Sine wave with noise.

The peak value of the noise is added to that of the sinusoidal voltage

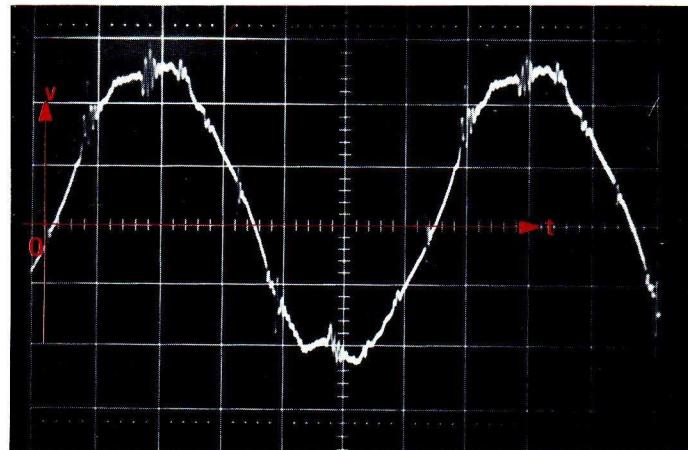


Fig. 7.5. Spikes on mains voltage

Some function generators on the market produce a sine wave which is shaped from a triangular wave by a diode-resistor network. If not compensated for, a small residue of the triangle appears just at the peaks and troughs of the sine wave (see fig. 7.6). Although the distortion factor of the sine wave generated can be specified very low, a peak-responding AC voltmeter will react to this residual triangle, resulting again in a faulty display. It goes without saying that the average-responding AC voltmeter will hardly react to this distortion.

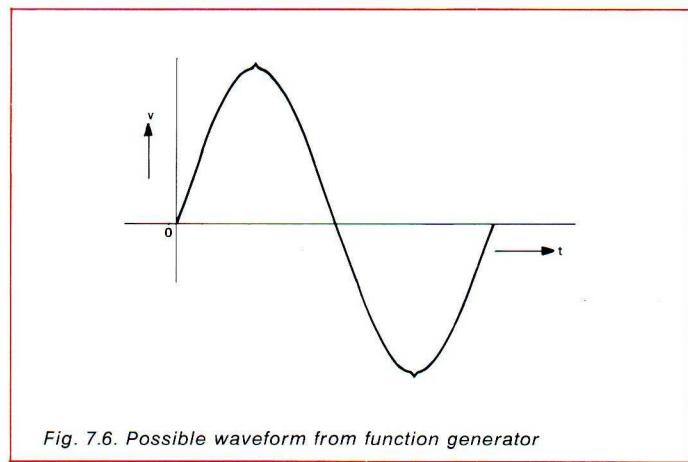


Fig. 7.6. Possible waveform from function generator

With harmonic distortion, the read-out of an average-responding and a peak-responding voltmeter can vary quite widely, depending on the circumstances. Figure 7.7 shows a distorted sine wave, composed of a fundamental plus its third harmonic with a relative amplitude of 10%. The two components are shown separately in figure 7.8. Given the phase shift shown in figure 7.8, it will be clear that an average-responding voltmeter will display the value of the fundamental only, since the third harmonic averages out. A peak-responding voltmeter will give a slightly higher value than the amplitude of the fundamental.

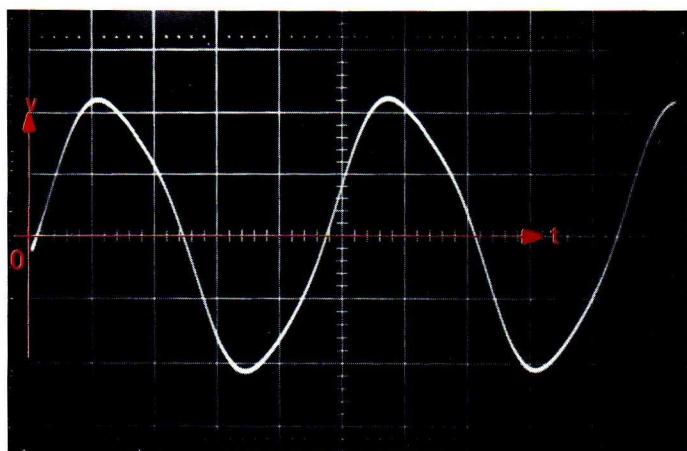


Fig. 7.7. Distorted sine wave with third harmonic

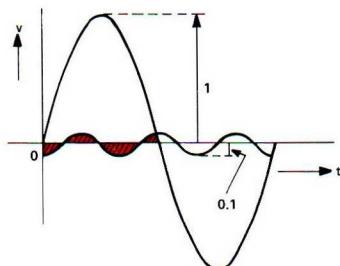


Fig. 7.8. Components of distorted sine wave of Fig. 7.7

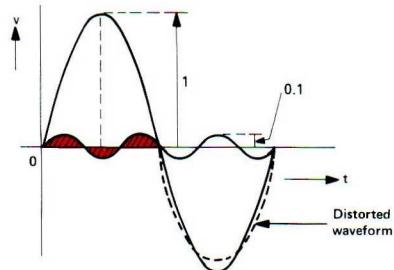


Fig. 7.9. Fundamental + 10% third harmonic; 3rd harmonic has a 90° phase shift with respect to that of Fig. 7.8

On the other hand, if the phase shift is as given in figure 7.9, it will be clear that this distortion of the waveform will lead to large errors in the reading. The average-responding voltmeter will give about 3% more than the fundamental and the peak-responding meter about 10% less. So, depending on the waveform and the type of meter used, we can expect errors of anything up to $\pm 10\%$ for the same amount of distortion. We shall see later on that the RMS voltmeter gives the correct reading in both cases.

The meaning of RMS

Definition:

The EFFECTIVE or RMS VALUE of a voltage of any waveform is defined as being the value of a DC voltage which will produce the same amount of energy in the same time.

- As already stated in the caption to figure 7.1., it is known that the RMS value of a sinusoidal voltage is $\frac{1}{\sqrt{2}}$ times

the peak value or $\frac{1}{2\sqrt{2}}$ times the peak-to-peak value.

- We all know that any periodic waveform can be expressed as the sum of an (infinite) series of sine waves, i.e. a fundamental with the same period as the waveform plus a series of higher harmonics. For example the waveform of figure 7.10 can be expressed as

$$v = v_1 \sin \omega t + v_2 \sin 2\omega t + v_3 \sin 3\omega t + \dots$$

Now let us denote the RMS values of the successive harmonics by V_1 , V_2 , V_3 , etc.

- We all know also that the power P developed in a resistor R by a DC voltage V_{DC} is given by

$$P = \frac{V_{DC}^2}{R}$$

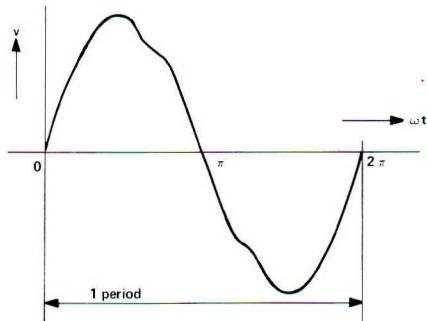


Fig. 7.10. Periodic waveform

- By definition, the complex waveform of figure 7.10 will develop a total power in R determined by its value:

$$P = \frac{V_{\text{RMS}}^2}{R}$$

If we resolve the waveform of figure 7.10 into its harmonics, then the RMS value of each harmonic will contribute its power to R, so the total power can also be expressed as

$$P = \frac{V_1^2}{R} + \frac{V_2^2}{R} + \frac{V_3^2}{R} + \dots$$

$$P = \frac{V_1^2 + V_2^2 + V_3^2 + \dots}{R}$$

As these two expressions must be identical, we may write

$$V_{\text{RMS}}^2 = V_1^2 + V_2^2 + V_3^2 + \dots$$

or

$$V_{\text{RMS}} = \sqrt{V_1^2 + V_2^2 + V_3^2 + \dots}$$

If v were to contain a **DC component** too, its power has to be added as well, so that the total power in this case becomes

$$P = \frac{V_{\text{DC}}^2}{R} + \frac{V_1^2 + V_2^2 + V_3^2 + \dots}{R}$$

whence

$$V_{\text{RMS}} = \sqrt{V_{\text{DC}}^2 + V_1^2 + V_2^2 + V_3^2 + \dots}$$

or

$$V_{\text{RMS}} = \sqrt{V_{\text{DC}}^2 + V_1^2 + V_2^2 + V_3^2 + \dots}$$

For a more detailed mathematical treatment see the Appendix.

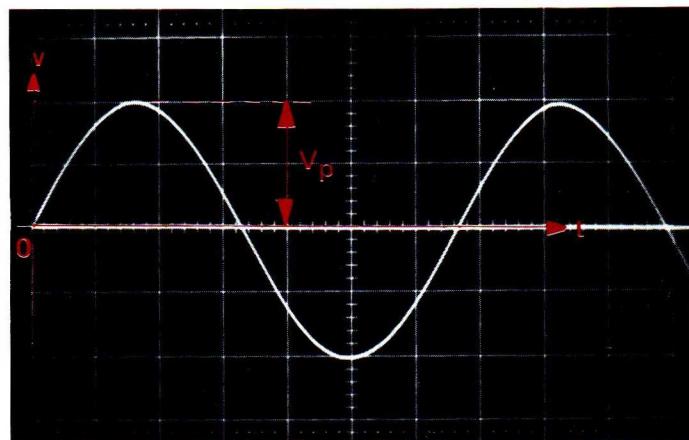


Fig. 7.11. Sinusoidal voltage

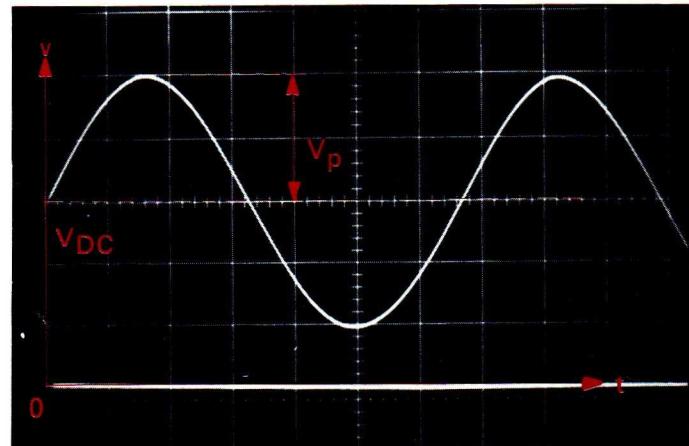


Fig. 7.12. Sine wave superimposed on DC level

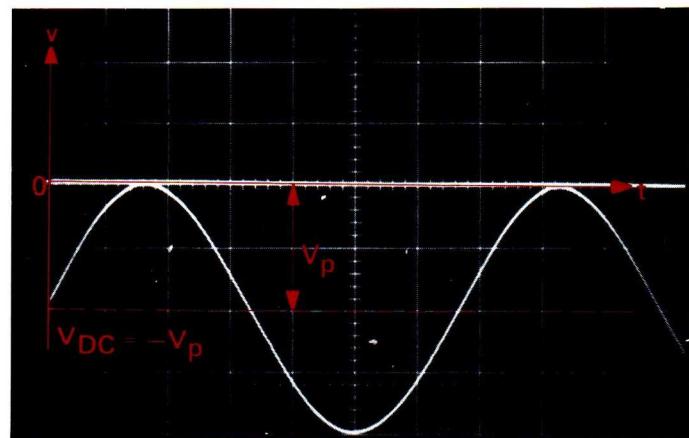


Fig. 7.13. Sine wave superimposed on DC level $V_{\text{DC}} = -V_p$

Examples

The RMS value of the sinusoidal voltage of figure 7.11 is

$$V_{\text{RMS}} = \frac{V_p}{\sqrt{2}}$$

or

$$V_{\text{RMS}}^2 = \frac{V_p^2}{2}$$

If, however, the sinusoidal voltage is superimposed on a DC voltage V_{DC} (figure 7.12) then the RMS value becomes

$$V_{\text{RMS}}^2 = V_{\text{DC}}^2 + \frac{V_p^2}{2}$$

or

$$V_{\text{RMS}} = \sqrt{V_{\text{DC}}^2 + \frac{1}{2}V_p^2}$$

Note that V_{DC} can be either positive or negative for the same value. If, for instance, $V_{\text{DC}} = -V_p$ (fig. 7.13) the RMS value becomes:

$$V_{\text{RMS}}^2 = (-V_p)^2 + \frac{V_p^2}{2}$$

or

$$V_{\text{RMS}} = V_p \sqrt{3/2} \approx 1.22 V_p$$

This last example makes it clear that we cannot just add the DC component and the RMS value of the AC component to get the overall RMS value.

The RMS detector in the PM 2527 digital multimeter

The operation of the RMS converter in our PM 2527 is based on the principle mentioned in the previous section: the power developed by a DC voltage is compared with that developed by the input voltage.

Figure 7.14 gives the circuit diagram, as designed in our Research Laboratory at Eindhoven and further developed in the Almelo factory.

The converter consists of two identical integrated circuits, IC_1 and IC_2 , each consisting of a resistor and a transistor, R_1 - TS_1 and R_2 - TS_2 respectively. Resistor R_1 is heated up by the input voltage V_{in} while R_2 is heated up by the output voltage of differential amplifier A. In each IC the temperature rise is measured by the transistor and the two temperatures are equalized by the control function of amplifier A. If transistor TS_1 is heated up by R_1 the current through it increases and the voltage at the inverting input of amplifier A decreases. This will result in a higher output voltage, causing R_2 to be heated up. This in its turn causes an increase in the current through TS_2 , reducing the voltage at the non-inverting input of amplifier A and hence the output voltage. When balance is reached, the

DC voltage from A equals the RMS value of the input voltage, because R_1 is equal to R_2 .

To ensure that the IC's are as nearly identical as possible, two adjacent ones from the same slice are selected during manufacture. They are not separated until they are mounted on a small glass plate that serves for thermal insulation (see figure 7.15).

The IC's are subsequently put into one housing to equalize the influence of the ambient temperature.

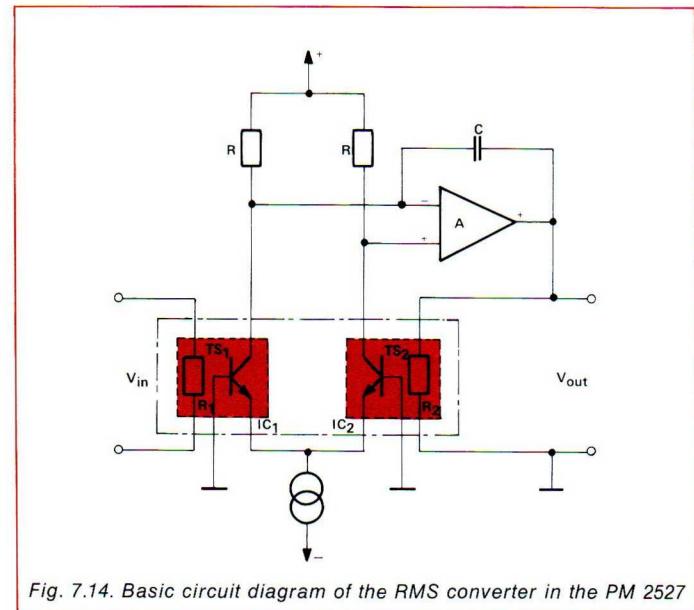


Fig. 7.14. Basic circuit diagram of the RMS converter in the PM 2527

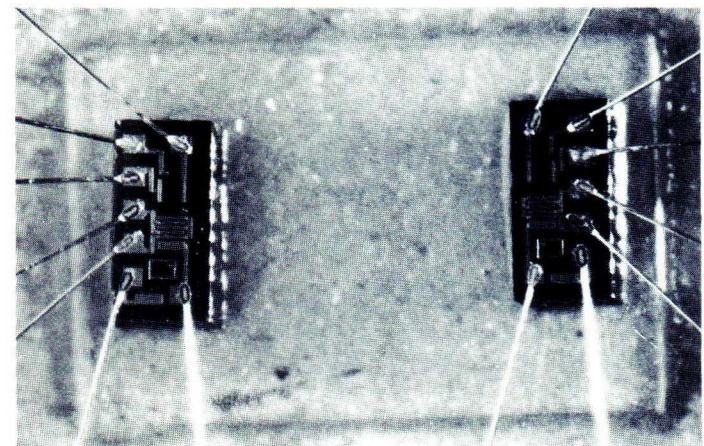


Fig. 7.15. Photograph of the RMS converter in the PM 2527. The glass plate measures about 2x3 mm, the chips about 1x0.5 mm

RMS specifications

AC-DC coupling

We see from figure 7.14 that the input voltage can be DC, while the output voltage always is DC. Basically, this circuit is thus a true RMS-to-DC converter. If, however, the converter is preceded by an AC-coupled preamplifier the system as a whole is AC-coupled (see figure 7.16).

• Lower frequency limit

Referring to the description of the operation of the RMS converter given above we see that at a certain instant equilibrium is reached so that $V_{\text{out}} = V_{\text{RMS}}$. Because the heating process in the IC's takes some time, as all heating processes do, equilibrium is only reached after about 1 sec. If now a frequency of 0.1 Hz is applied to the input, the system will continuously vary V_{out} in an attempt to make it equal to the input voltage, and no stable output will be obtained. This is why the lower frequency is set at 30 Hz: a variation with this frequency is too fast for the thermal system, so a stable output is obtained.

Note that this will also be the case with a DC-coupled preamplifier; strictly speaking, therefore, the specification should then read:

Frequency range: 30 Hz–100 kHz + DC.

• Upper frequency limit

This limit is set by the preamplifier and IC₁ in figure 7.14, i.e. the input system as a whole. The upper frequency limit becomes important when we consider the number of harmonics needed to approximate to an arbitrary, periodical waveform within the limits set by the specified accuracy. If for example we want to measure the RMS value of a square-wave of ideal shape, i.e. with rise and fall times equal to zero (figure 7.17) we need a series of 395 harmonics before the RMS sum of all the harmonics is equal to 99.9% of the theoretical value. So, if the frequency of the square-wave is 50 Hz (period 20 ms), we need a bandwidth of 20 kHz for an intrinsic error of 0.1%...

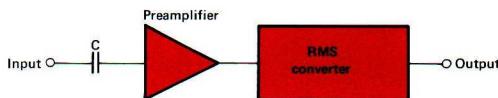


Fig. 7.16. AC-coupled system

But this is theory!

In practice, the rise and fall times are finite so that about 20 harmonics are sufficient for a proper approximation in 95% of the cases. In figure 7.17 this would mean that a 10–90% rise time of 2% of the period time can be approximated by 19 harmonics within 0.1%.

Also thyristor power-control circuitry generates waveforms with extremely short rise and fall times, but fortunately stray capacitances in the surrounding circuitry or built-in filters suppress all the higher harmonics. Moreover the basic frequency is generally 50 (60) Hz or 100 (120) Hz so that here again the specified 100 kHz bandwidth of the PM 2527 will be more than sufficient.

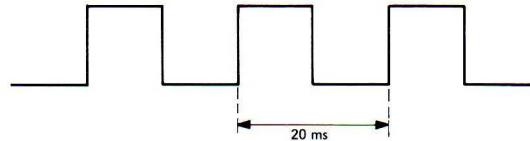


Fig. 7.17. Ideal square-wave

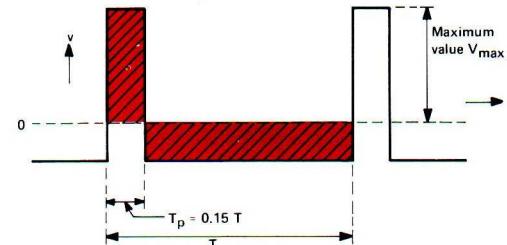


Fig. 7.18. Pulse with crest factor of 2.4

Accuracy

This depends on linear operation of the RMS converter. If the non-linearity of this converter is 0.1% then together with the surrounding electronics it will give a specified accuracy of 0.2% for the lower frequencies (up to 40 kHz for the PM 2527). At higher frequencies the stray capacitances of the preamplifier and attenuator become important and the specified accuracy will decrease slightly (0.5% up to 100 kHz for the PM 2527).

It is for this reason that the display of the PM 2527 shows only 4 digits for AC measurements since the 5th digit is meaningless at this accuracy.

Crest factor (CF)

The crest factor of an alternating voltage is defined as the

ratio of the maximum value to the RMS value, $\frac{V_{\max}}{V_{\text{RMS}}}$

For example, if the waveform is a pulse as in figure 7.18, the RMS value can be calculated (see Appendix) to be:

$$V_{\text{RMS}} = V_{\max} \sqrt{\frac{T_p}{T - T_p}}$$

Since $T_p = 0.15 T$ (see fig. 7.18) we find for the crest factor

$$CF = \frac{V_{\max}}{V_{\text{RMS}}} = \frac{\sqrt{T - 0.15T}}{0.15T} \approx 2.4$$

Note that this pulse is averaged around zero (AC coupling). If the same pulse were to be clamped at zero (fig. 7.19), then it can be shown that

$$V_{\text{RMS}} = V_{\max}^1 \sqrt{\frac{T_p}{T}}$$

and the crest factor CF becomes

$$CF = \frac{V_{\max}^1}{V_{\text{RMS}}} = \sqrt{\frac{T}{T_p}} \approx 2.6$$

This example shows the importance of a proper choice between DC and AC coupling, and explains why the definition of the crest factor is based on the **maximum** value of the voltage. As far as the instrument is concerned, this specification is determined by the ratio of the dynamic range of the input amplifier (which depends on its supply voltage) to the RMS voltage at the input of the converter giving maximum output. If the amplitude of the input voltage giving maximum converter output exceeds the dynamic range of the input amplifier, the latter clips off the peaks of the input voltage and only part of the (distorted) input voltage is measured.

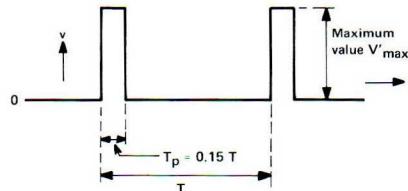


Fig. 7.19. Pulse clamped at zero

NOTE

The crest factor of 2.4 for the PM 2527 is specified for a full-range display of 1999. Some manufacturers specify a full range of 999 with 100% overrange up to 1999. If a crest factor of 5 at full range is specified under these conditions, this refers to a full range of 999. For a full-range display of 1999, one needs twice the input voltage, so with the same input amplifier with the same dynamic range the crest factor will be reduced by a factor of two, i.e. to 2.5 in the present example. This must be borne in mind when reading the specifications.

RMS and TRUE RMS

As it will be clear by now, the DC component of a signal also contributes to the RMS value, so only in those cases where we have DC coupling can we really measure the true RMS value of the complete signal.

An RMS voltmeter which only has an AC-coupled input is thus not a TRUE RMS meter – since we can never determine with this instrument whether a DC component is present in the input signal or not.

Two simple examples

1-CREST FACTOR OF SINE WAVE

A sinusoidal voltage of amplitude V_p ($= V_{\max}$) and hence an RMS value of $\frac{V_p}{\sqrt{2}}$ has by definition a crest factor CF of

$$CF = \frac{V_p}{V_{\text{RMS}}} = \sqrt{2} = 1.4 \text{ (see figure 7.20)}$$

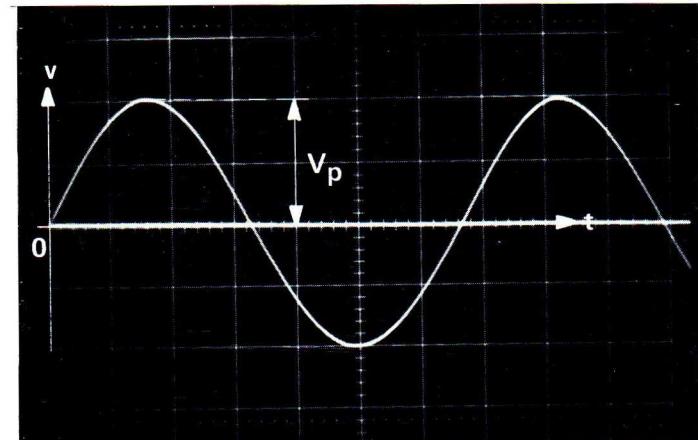


Fig. 7.20. The crest factor of a sine wave is 1.4

2-CREST FACTOR OF SQUARE WAVE

Since the RMS value is obtained by summing the squares of the component voltages, it makes no distinction between positive and negative values; it will thus be clear that the RMS value of a pure 1/2 T square pulse equals the peak value V_p (see figure 7.21), so by definition the crest factor

$$CF = \frac{V_p}{V_{RMS}} = 1.$$

Note that here the average is zero (AC coupling). If the square wave is clamped to zero, this is equivalent to adding a DC voltage of V_p to it (figure 7.22). In this case $V_{RMS} = 1/2\sqrt{2} \cdot V_p$ and the crest factor CF becomes

$$CF = \frac{V_p}{V_{RMS}} = \sqrt{2} = 1.4$$

Some more practical examples will be dealt with below

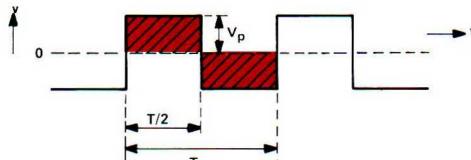


Fig. 7.21. The crest factor of a square wave is 1

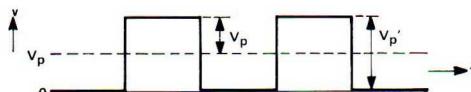


Fig. 7.22. A clamped square wave has a crest factor of 1.4

$$CF = \sqrt{\frac{1-\alpha}{\alpha}}$$

and for the clamped pulse

$$CF = \sqrt{\frac{1}{\alpha}}$$

with α = duty cycle.

These functions are plotted in Fig. 7.23.

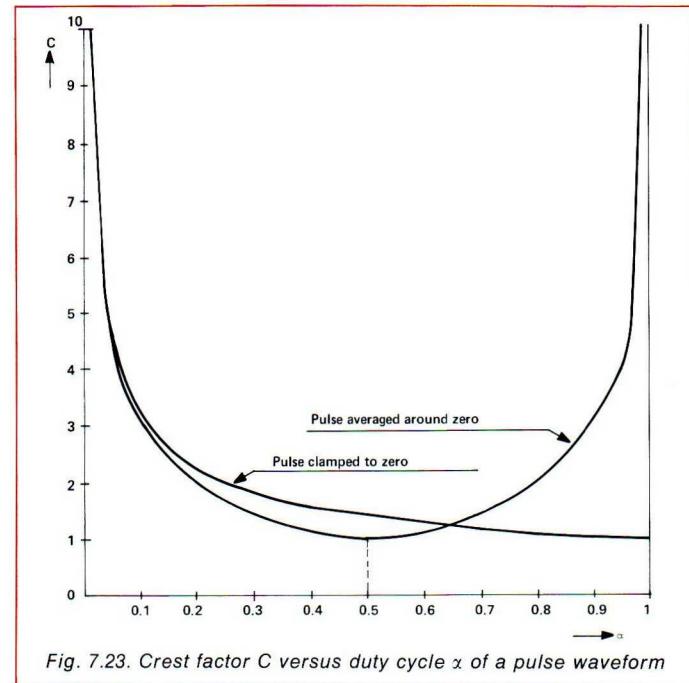


Fig. 7.23. Crest factor C versus duty cycle α of a pulse waveform

RMS measurements in practice

Crest factor of pulses

In figure 7.19 it was shown that for a clamped pulse to zero the crest factor CF is

$$CF = \sqrt{\frac{T}{T_p}}$$

and for an averaged pulse around zero

$$CF = \sqrt{\frac{T-T_p}{T_p}}$$

(see also Appendix)

If we consider the above expressions we can plot the crest factor CF versus the duty cycle α of a pulse waveform. For the pulse averaged around zero we get



Fig. 7.24. 1 kHz sine wave with 6% distortion

Distorted sine wave

Figure 7.24 shows the experimental set-up used for measuring a distorted sine wave. The measured distortion was 6% and although the trace looks like a sine wave to the eye, the difference in read-out speaks for itself.

From this example it will be clear that only the RMS meter gives the right answer when measurements have to be made on distorted sine waves.

We saw in connection with figures 7.7 and 7.8, that 10% harmonic distortion can give a measuring error of about 3% with an average-responding AC voltmeter. Figure 7.25 plots some measuring errors for the 3rd, 5th and 7th harmonics over a wider range of distortion.

Switched sine wave

Switched sine waves are often encountered in power control; and since power is involved, RMS measurements are very important. Here again, (see figure 7.26) the average-responding AC meter will give rather large errors. It would take us beyond the scope of this book to give an analytical calculation of the error, but the results of such a calculation are plotted in fig. 7.27. We see that for $\alpha = 0.1$ (10% cut off) the error is already about 2%. The photograph of fig. 7.28 gives an actual set-up used for measurements on switched sine waves: a 3-phase primary-controlled high voltage generator is being tested before delivery.

Another interesting question is how much of the sine wave we can switch off before we reach the limiting crest factor (2.4 for the PM 2527).

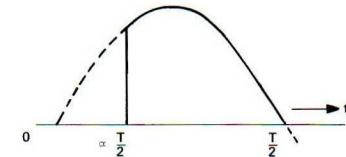


Fig. 7.26. Switched sine wave

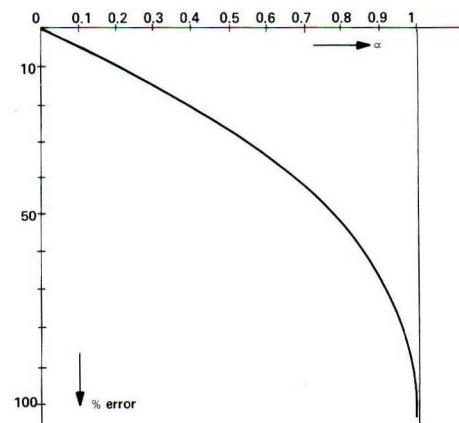


Fig. 7.27. Percentage error indicated by an average-responding voltmeter as function of the switched-off part α of a sine wave

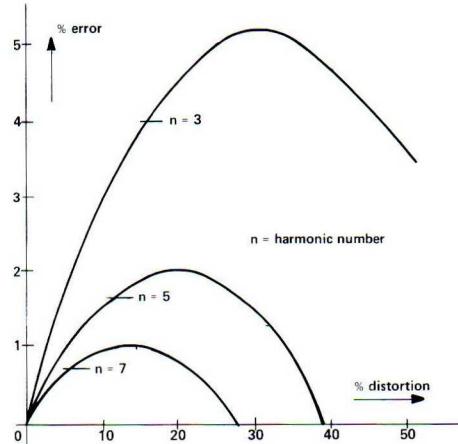


Fig. 7.25. Error made in measuring distorted sine wave with an RMS calibrated average-responding AC meter

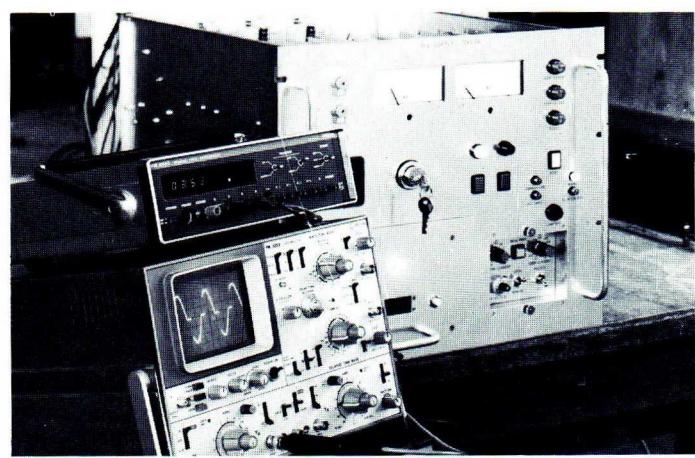


Fig. 7.28. Practical set-up for RMS measurements on a power-control system

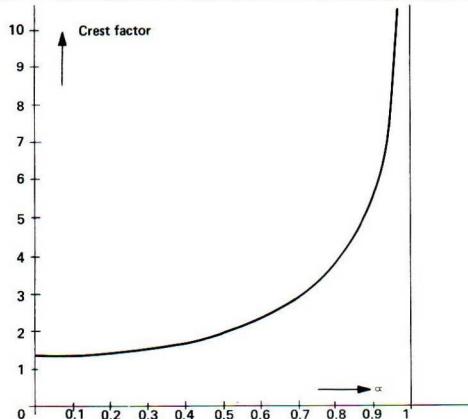


Fig. 7.29. Crest factor as function of α = switched-off part of the sine wave.

Figure 7.29 shows the plot of the crest factor as a function of α for this case. We see from this that we can switch off about 60% ($\approx 110^\circ$) of the sine wave before the crest factor of the rest reaches 2.4.

Fig. 7.30 gives a visual impression of what this means.

Full-wave rectified sine wave

If we remember the theory of RMS given above, it will be clear that the RMS value of a full-wave rectified sine wave equals that of the sine wave itself. So if the amplitude is 1 (fig. 7.31), the RMS value is $1/\sqrt{2} = 0.707$. This is the TRUE RMS value.

An AC-coupled instrument will subtract the DC content of this waveform, according to the rule

$$V_{\text{TRUE RMS}} = \sqrt{V_{\text{DC}}^2 + V_{\text{AC RMS}}^2}$$

or

$$V_{\text{AC RMS}} = \sqrt{V_{\text{TRUE RMS}}^2 - V_{\text{DC}}^2}$$

The average value is $2/\pi$ (see fig. 7.1.), so

$$V_{\text{AC RMS}} = \sqrt{\frac{1}{2} - 4/\pi^2}$$

It follows from this that

$$\frac{V_{\text{TRUE RMS}}}{V_{\text{AC RMS}}} = \sqrt{\frac{1}{2}/(\frac{1}{2} - 4/\pi^2)} = 2.29760$$

Thus we can state to within an accuracy of 0.1% that $V_{\text{TRUE RMS}} = 2.3 V_{\text{AC RMS}}$

For the AC-coupled PM 2527 this means for full-wave rectified sine waves we can simply multiply the measured result by a factor of 2.3 in order to know the TRUE RMS value and the result remains WITHIN THE INSTRUMENT'S ACCURACY SPECIFICATIONS.

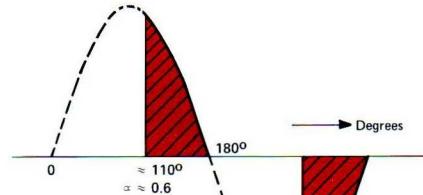


Fig. 7.30. Switched sine wave the crest factor of which is 2.4.

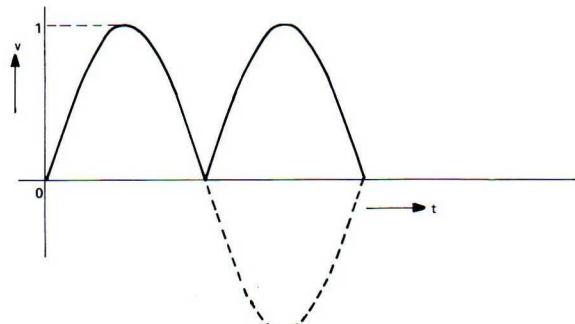


Fig. 7.31. For a full-wave rectified sine wave, $V_{\text{TRUE RMS}} = 2.3 \times V_{\text{AC RMS}}$

Noise measurements

It will be clear from what has been said above that an RMS voltmeter is pre-eminently suited for the measurement of noise voltages. The PM 2527 is even better suited than an RMS meter: single-scan measurements can be performed by pushing the START tumbler switch to the MAN (manual) position. When the button is released it flips back to the EXT position. In this way a stable read-out is obtained, while in the AUT position (automatic) the read-out will vary with the noise.

Figure 7.32 also illustrates the rule of the thumb that the peak-to-peak value of a noise voltage is about 7 to 8 times the RMS value. The noise voltage covers practically the entire screen, i.e. the peak-to-peak noise voltage is about 7–8 V (oscilloscope setting 1 V/div.) while the reading displayed on the PM 2527 is about 1 V.

Appendix

Mathematical aspects of RMS

If we apply a voltage to a resistor R , the resistor will be heated up, as we all know. The amount of energy E involved was first measured by JOULE, who found the relation:

$$\Delta E = \frac{v^2}{R} \Delta t$$

For the discussion of voltages which vary with time, we choose the time Δt short; for an infinitesimally short time dt the amount of energy dE may be written:

$$dE = \frac{v^2}{R} dt$$

The total amount of energy developed as heat in the resistor between $t=0$ and $t=T$ is thus:

$$E = \frac{1}{R} \int_0^T v^2 dt$$

For a DC voltage V_{DC} this expression leads to

$$E = \frac{V_{DC}^2}{R} T$$

If we now define the EFFECTIVE or RMS VALUE of a voltage v as being the value of the DC voltage which would produce the same amount of heat in a certain time, then it follows from the last two expressions that

$$\frac{1}{R} \int_0^T v^2 dt = \frac{V_{DC}^2}{R} T$$

Now by definition $V_{DC} = V_{RMS}$, so that

$$\int_0^T v^2 dt = V_{RMS}^2 T$$

or

$$V_{RMS} = \sqrt{\frac{1}{T} \int_0^T v^2 dt}$$



Fig. 7.32. Measuring noise with the PM 2527. The oscilloscope is set to a vertical sensitivity of $1 \text{ V}/\text{div}$

Examples

- Sine wave (see figure 7.33 or figure 7.1):

$$v = V_p \sin \omega t \quad \omega T = 2\pi$$

$$V_{RMS} = \sqrt{\frac{V_p^2}{T} \int_0^T \sin^2 \omega t dt}$$

$$V_{RMS} = V_p \sqrt{\frac{1}{T} \int_0^T \left(\frac{1}{2} - \frac{1}{2} \cos 2\omega t\right) dt}$$

$$V_{RMS} = V_p \sqrt{\frac{1}{T} \left[\frac{1}{2}t - \frac{1}{2} \frac{1}{2\omega} \sin 2\omega t \right]_0^T}$$

$$V_{RMS} = V_p \sqrt{\frac{1}{T} \left[\frac{1}{2}T \right]}$$

$$V_{RMS} = \frac{V_p}{\sqrt{2}}$$

as stated in figure 7.1. for $V_p = 1$

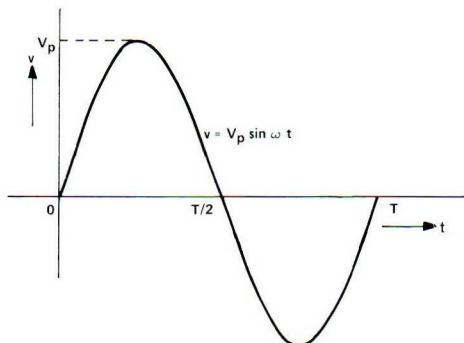


Fig. 7.33. Pure sine wave

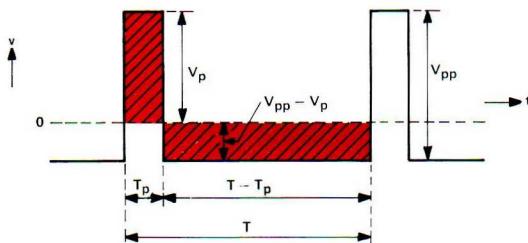


Fig. 7.34. Pulse wave of which the average value is zero

• Pulse (see figure 7.34):

For a waveform that is symmetrical around zero, we may write:

$$V_p T_p = (V_{pp} - V_p)(T - T_p)$$

or

$$V_{pp} = \frac{T}{T - T_p} V_p$$

Now the RMS value of v becomes

$$V_{RMS} = \sqrt{\frac{1}{T} \int_0^T v^2 dt}$$

$$V_{RMS} = \sqrt{\frac{1}{T} \left[\int_0^{T_p} V_p^2 dt + \int_{T_p}^T (V_{pp} - V_p)^2 dt \right]}$$

$$V_{RMS} = \sqrt{\frac{1}{T} \left[V_p^2 t \Big|_0^{T_p} + (V_{pp} - V_p)^2 t \Big|_{T_p}^T \right]}$$

$$V_{RMS} = \sqrt{\frac{1}{T} [V_p^2 T_p + (V_{pp} - V_p)^2 (T - T_p)]}$$

$$\text{with } V_{pp} = \frac{T}{T - T_p} V_p$$

$$V_{RMS} = V_p \sqrt{\frac{T_p}{T - T_p}}$$

as stated in connection with figure 7.18 for $V_{max} = V_p$

Questions

Q.7.1. The average value of a sine wave over a whole period is:

A $1/\sqrt{2}$

B 0

C $2/\pi$

| | |
|---|--|
| A | |
| B | |
| C | |

Q.7.2. The circuit of fig. 7.35 is:

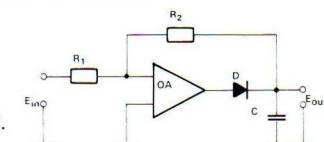


Fig. 7.35.

A a fullwave rectifier

B a halfwave rectifier

C an averaging circuit

| | |
|---|--|
| A | |
| B | |
| C | |

Q.7.3. For a distorted sine wave the components of which are given in fig. 7.36, an average-responding RMS voltmeter will display a voltage which is:

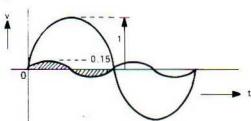


Fig. 7.36.

A too high
B correct
C too low

| | |
|---|--|
| A | |
| B | |
| C | |

Q.7.4. A peak-responding voltmeter receiving an input waveform as shown in fig. 7.36 will display a voltage which is:

A too high
B correct
C too low

| | |
|---|--|
| A | |
| B | |
| C | |

Q.7.5. The theoretically true RMS value of a sine wave superimposed on a negative DC voltage is:

A higher than
B the same as
C lower than
the RMS voltage of the sine wave alone

| | |
|---|--|
| A | |
| B | |
| C | |

Q.7.6. The true RMS value of a sinewave voltage of amplitude $V_p = 4\sqrt{2}$ V superimposed on a DC voltage $V_{DC} = 3$ V is:

A 5V
B $5\sqrt{2}$ V
C $7\sqrt{2}$ V

| | |
|---|--|
| A | |
| B | |
| C | |

Q.7.7. Basically, the RMS converter part in the PM 2527 is:

A a true RMS-to-DC converter
B an AC-coupled RMS detector
C an averaging circuit

| | |
|---|--|
| A | |
| B | |
| C | |

Q.7.8. An AC-coupled RMS voltmeter displays:

A the DC component of the signal
B the true RMS value
C the RMS value

| | |
|---|--|
| A | |
| B | |
| C | |

Q.7.9. If we connect an AC-coupled amplifier in front of an RMS converter the crest-factor specification will be:

A higher than
B the same as
C lower than
that of the RMS converter alone.

| | |
|---|--|
| A | |
| B | |
| C | |

Q.7.10. The RMS value of a square-wave voltage averaged around zero with a peak value of $V_p = 1$ V (see fig. 7.37) is:

A $1\sqrt{2}$
B 1
C $\sqrt{2}$

| | |
|---|--|
| A | |
| B | |
| C | |

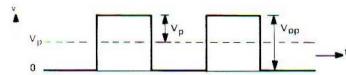


Fig. 7.37.

Q.7.11. If only the amplitude of a pulsed waveform decreases, then the crest-factor:

A decreases
B remains the same
C increases

| | |
|---|--|
| A | |
| B | |
| C | |

Q.7.12. If the duty cycle of pulse clamped to zero decreases, then the crest-factor of the pulse:

A decreases
B remains the same
C increases

| | |
|---|--|
| A | |
| B | |
| C | |

Q.7.13. If we decrease the duty cycle α of a pulse clamped to zero from nearly 100% to 25% of its initial value and we measure still in the same RMS voltmeter range, the read-out will be multiplied by:

A 2
B $\frac{1}{2}$
C $\frac{1}{4}$

| | |
|---|--|
| A | |
| B | |
| C | |

Q.7.14. Let us suppose that a given RMS voltmeter has a 100% over-range facility, permitting readings up to 200% of the specified full scale. Its crest-factor is specified at full range (100%). With a 200% read-out, the maximum permissible crest-factor of the signal is the specified crest-factor multiplied by:

A 2
B $\sqrt{2}$
C $\frac{1}{2}$

| | |
|---|--|
| A | |
| B | |
| C | |

Q.7.15. The theoretical TRUE RMS value of a full-wave-rectified sine wave is:

A higher than
B the same as
C lower than
that of the original sine wave itself

| | |
|---|--|
| A | |
| B | |
| C | |

Q.7.16. The AC-coupled digital RMS multimeter PM 2527 displays the RMS value of a full-wave-rectified sine wave

A too high
B correct
C too low
compared with the theoretical value

| | |
|---|--|
| A | |
| B | |
| C | |

Q.7.17. For a noise voltage we may write: $V_{p-p} : V_{RMS} = 8$. For a correct reading on the PM 2527 with a crest-factor of 2.4, the indicated value should not exceed the following percentage of full range:

A 30%
B 60%
C 66.7%

| | |
|---|--|
| A | |
| B | |
| C | |

Answers to questions

Chapter 6

A.6.1 C

A.6.2 C

A.6.3 B

A.6.4 A

A.6.5 B $X_c = \frac{1}{2\pi \times 2.6 \times 10^{-9} \times 60} \approx 1 \text{ M}\Omega$

so

$$V_{zt} = \frac{1}{1 + 1000} E_{cm} \approx \frac{E_{cm}}{1000}$$

$$\text{CMRR} = 20 \log \frac{E_{cm}}{E_{cm \text{ measured}}} \text{ dB}$$
$$= 60 \text{ dB}$$

(see also text referring to fig. 6.15).

Chapter 1

A.1.1 B

A.1.2 A and B

A.1.3 B

A.1.4 A and C

A.1.5 A

Chapter 2

A.2.1 B

A.2.2 C

A.2.3 A true
B true
C not true

A.2.4 A and C

A.2.5 B and C

Chapter 3

A.3.1 C

A.3.2 A and C

A.3.3 B and C

A.3.4 B

Chapter 4

A.4.1 C

A.4.2 A

Chapter 5

A.5.1 B 0.05% of reading is 0.11 V and 0.02% of range is 0.40 V, together 0.51 V or 0.23% of 220 V.

A.5.2 C

A.5.3 B

A.5.4 A true
B not true
C not true

Chapter 7

A.7.1 B positive and negative parts are equal.

A.7.2 B one diode, no series resistor.

A.7.3 B second harmonic averages out.

A.7.4 A combining the two components gives an amplitude > 1 for part of the waveform.

A.7.5 A the DC voltage contributes to the power as the square of its value.

A.7.6 A for $V_p = 4\sqrt{2}$ volt, $V_{RMS} = 4$ V; so V_{RMS} for the combined signal is $\sqrt{3^2 + 4^2} = 5$

A.7.7 A both DC and AC signals heat up the input resistor.

A.7.8 C the DC component is not displayed.

A.7.9 A the pulses are not clamped, but averaged around zero.

A.7.10 B RMS value = peak value.

A.7.11 B crest value depends only on duty cycle.

A.7.12 C crest factor is $1/\sqrt{\alpha}$

A.7.13 B RMS value is peak value times $\sqrt{\alpha}$

A.7.14 C for the same dynamic range of the input amplifier, the read-out can only be increased by increasing the duty cycle, thus decreasing the crest factor.

A.7.15 B voltages are squared in RMS calculations.

A.7.16 C DC power is subtracted.

A.7.17 B $V_p = \frac{1}{2}V_{pp}$ so crest factor is 4, and $2.4/4 = 0.6$.

